WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

AUGUST 1956

Vol. 33 No. 8 · THREE SHILLINGS AND SIXPENCE

See AVO Instruments at the
RADIO SHOW STAND No. 3

MODEL 7 50 RANGE
Universal AVOMETER

Fifty ranges...

in one instrument

THE wide scope of this multi-range AC/DC measuring instrument, coupled with its unfailing reliability, simplicity of use and high degree of accuracy, renders it invaluable wherever electrical equipment has to be maintained in constant, trouble-free operation.

CURRENT	VOLTAGE
1mA D.C. only	50mV D.C. only
2 „ „ „	100 „ „ „
5 „ A.C. & D.C.	500 „ „ „
10 „ „ „	1V „ „
50 „ „ „	5 „ A.C. & D.C.
100 „ „ „	10 „ „ „
500 „ „ „	50 „ „ „
1 A „ „	100 „ „ „
5 „ „ „	200 „ „ „
10 „ „ „	400 „ „ „
	500 „ „ „
	1,000 „ „ „

RESISTANCE

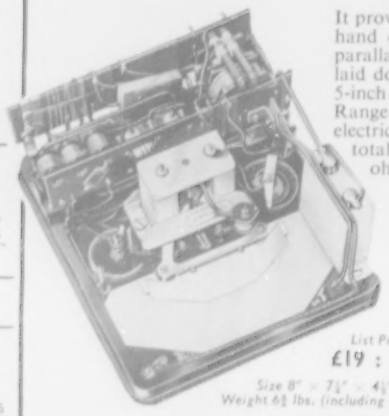
10,000 ohms, 100,000 „ } using internal 1½ volt cell
1 Megohm using internal 9 volt battery
10 Megohms, 40 „ } using external source of A.C. or D.C. voltage

CAPACITY 0·01—20 mFds

POWER AND DECIBELS

Impedance	Power	Decibels 0 50W
500 ohms	200 mW	—25 to + 6
5,000 ohms	2W	—15 to + 16
50,000 ohms	200 mW	—25 to + 6

Various accessories are available for extending the wide range of measurements.



It provides 50 ranges of readings on a 5-inch hand calibrated scale fitted with an anti-parallax mirror. Accuracy is within the limits laid down in Section 6 of B.S.S. 89/1954 for 5-inch scale industrial portable instruments. Range selection is effected by means of two electrically interlocked rotary switches. The total resistance of the meter is 500,000 ohms.

The instrument is self-contained, compact and portable, simple to operate, and is protected by an automatic cut-out against damage through inadvertent overload.

Power and Power Factor can be measured in A.C. circuits by means of an external accessory, the Universal AvoMeter Power Factor & Wattage Unit.

List Price
£19 : 10s.

Size 8" × 7¼" × 4¼"
Weight 6¾ lbs. (including leads)

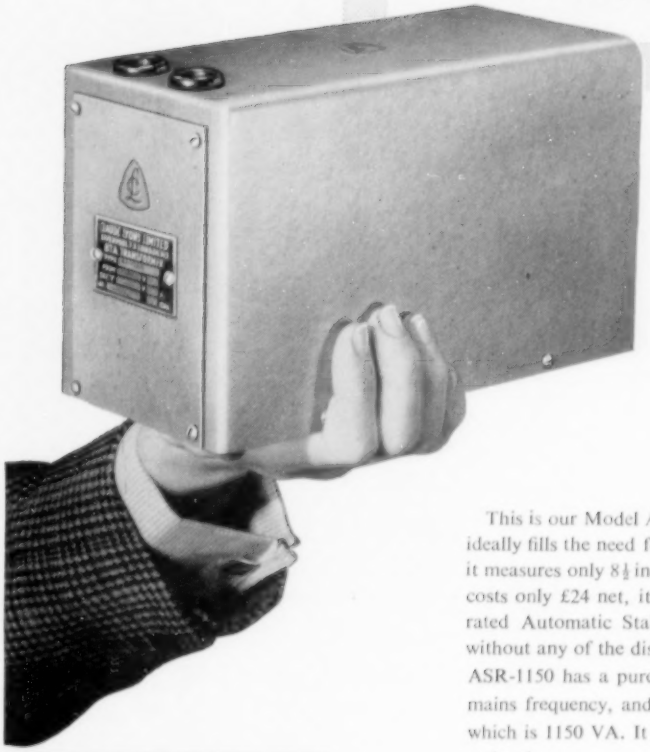
...you can depend on AVO

Illustrated Brochure available on request.

Sole Proprietors and Manufacturers:—
THE AUTOMATIC COIL WINDER & ELECTRICAL EQUIPMENT CO., LTD.
AVOCET HOUSE • 92-96 VAUXHALL BRIDGE ROAD • LONDON • S.W.1. *Telephone: VICtoria 3404 (9 lines)*

A7/9

A NEW-PRINCIPLE A-C AUTOMATIC VOLTAGE STABILISER



This is our Model ASR-1150 Automatic Voltage Step Regulator. It ideally fills the need for a cheap, small and light Stabiliser. Although it measures only 8½ in. × 4½ in. × 5 in., weighs as little as 11 lb., and costs only £24 net, it has a performance fully equal to any similarly rated Automatic Stabiliser of the resonated, saturated core type, without any of the disadvantages.

ASR-1150 has a pure output waveform, is unaffected by changes in mains frequency, and works equally well from no-load to full load, which is 1150 VA. It has a stabilised output at 230V unless otherwise ordered.

Many other Automatic Voltage Stabilisers are now manufactured by us, and all are available for immediate delivery. In some cases the constancy of output is as high as 0.15%. Models are available from 200 VA to 30 kVA, single phase. 3-Phase Stabilisers are also available. Prices are *extremely* competitive.

We can supply from stock all types of American tubes, condensers, valves, potentiometers, etc.

MEMO: If you are interested in infinitely-variable Transformers, do not forget the almost indispensable "VARIAC" (Reg'd. Trademark). Models are available from 170VA to 21kVA. Our Catalogue V-549 (3rd Edition) tells the whole story and will gladly be mailed free and post free, on request.

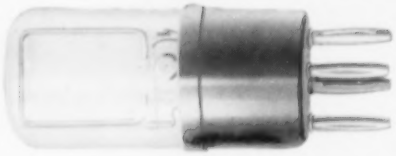
The *NEW "ASR-1150"* costs only *£24* net

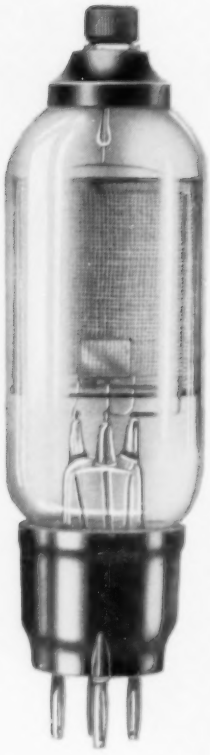
Complete information is obtainable from:

CLAUDE LYONS LTD., STABILISER DIVISION,

HEAD OFFICES & WORKS: 76 OLDHALL STREET, LIVERPOOL 3, LANCS.
SOUTHERN FACTORY: VALLEY WORKS, WARE ROAD, HODDESDON, HERTS.
(A10 main London/Cambridge Road, at Junction of A602)

PHOTO-ELECTRIC CELLS





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The new English Electric Valve brochure gives data of the widest range of valves for industry, communications, radar and broadcasting made in Great Britain. A copy of this publication will gladly be sent on request.

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Ferranti Valves for FM/AM Radio Receivers

The types illustrated are as follows: ECC85, ECH81, EF85, EABC80, EL84, EZ80, EM80.

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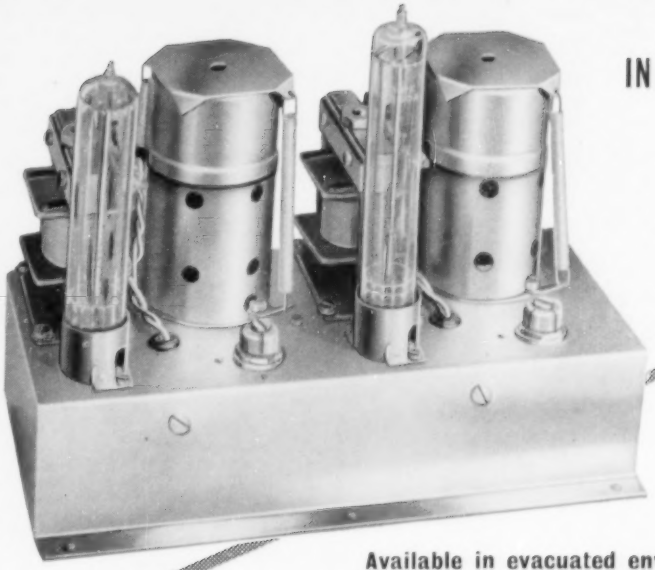
MARCONI LOW FREQUENCY CRYSTALS



IN THE RANGE 1.6-50 Kc/s

The illustration shows a 1.6 and 3 Kc/s Carrier Oscillator Unit for transmitting information to a remote bearing indicator as used in the Marconi V.H.F. Direction Finder Type AD200.

Available in evacuated envelopes with B7G base or with flexible leads for connection into the circuit

Among the many uses for this type of crystal are the following:—

SPEECH INVERSION OSCILLATORS

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CARRIER CHANNELLING EQUIPMENT

•

RANGE CALIBRATORS

•

TIMING APPLICATIONS ETC.

Flexure mode crystals are suitable for a wide variety of applications where an oscillator of good stability and small size is essential. They cover the frequency range 1.6 to 50 Kc/s with a point of zero temperature coefficient in the range 0°-40°C. The bar in XY flexure used between 1.6 and 12.0 Kc/s has the great advantage of a low temperature coefficient in the region of room temperature.

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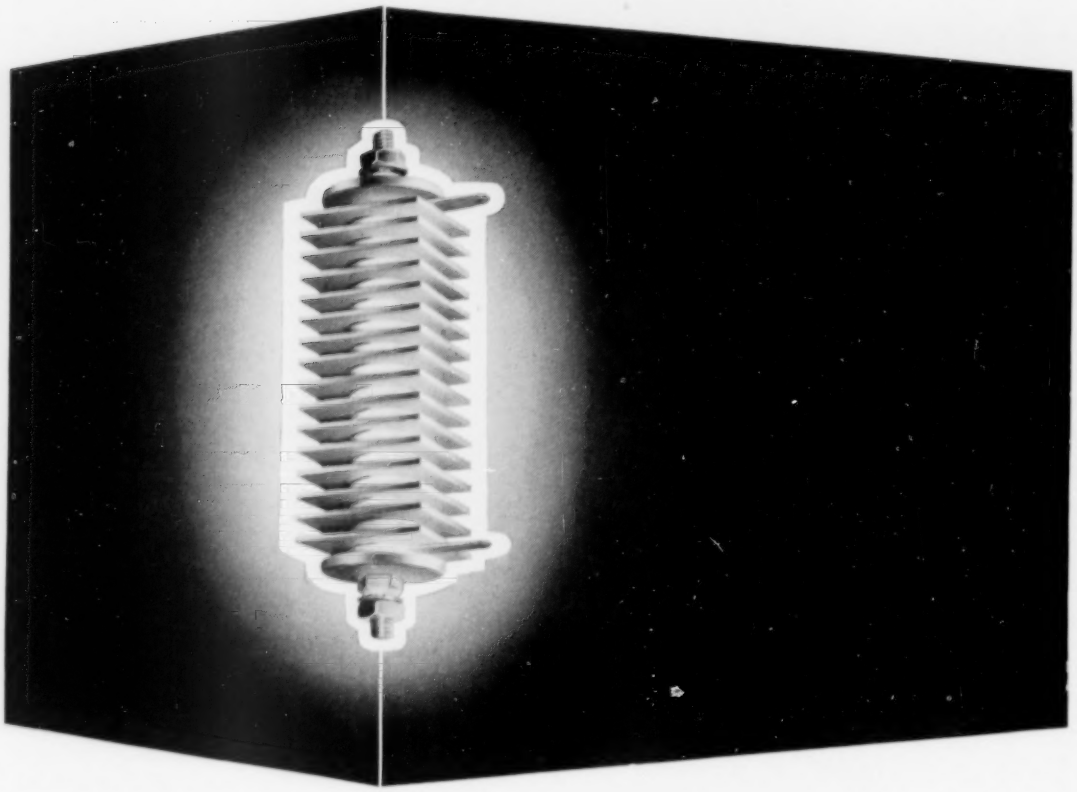

In the vanguard of progress

Since January, 1956, Salford Electrical Instruments Limited have been in regular production of Vacuum-deposited Selenium rectifiers employing Aluminium base plates. This is one of the most outstanding technical advances in selenium rectifier manufacture in this country in recent years. In conjunction with other improvements it results in reduced forward resistance and therefore increased current rating for the same size of elements, while at the same time maintaining the full inverse voltage rating. Further important advantages are an extended temperature range of operation, improved life and reduced weight.

A SIGNIFICANT CONTRIBUTION TO THE ADVANCE OF ELECTRICAL AND ELECTRONIC TECHNIQUES

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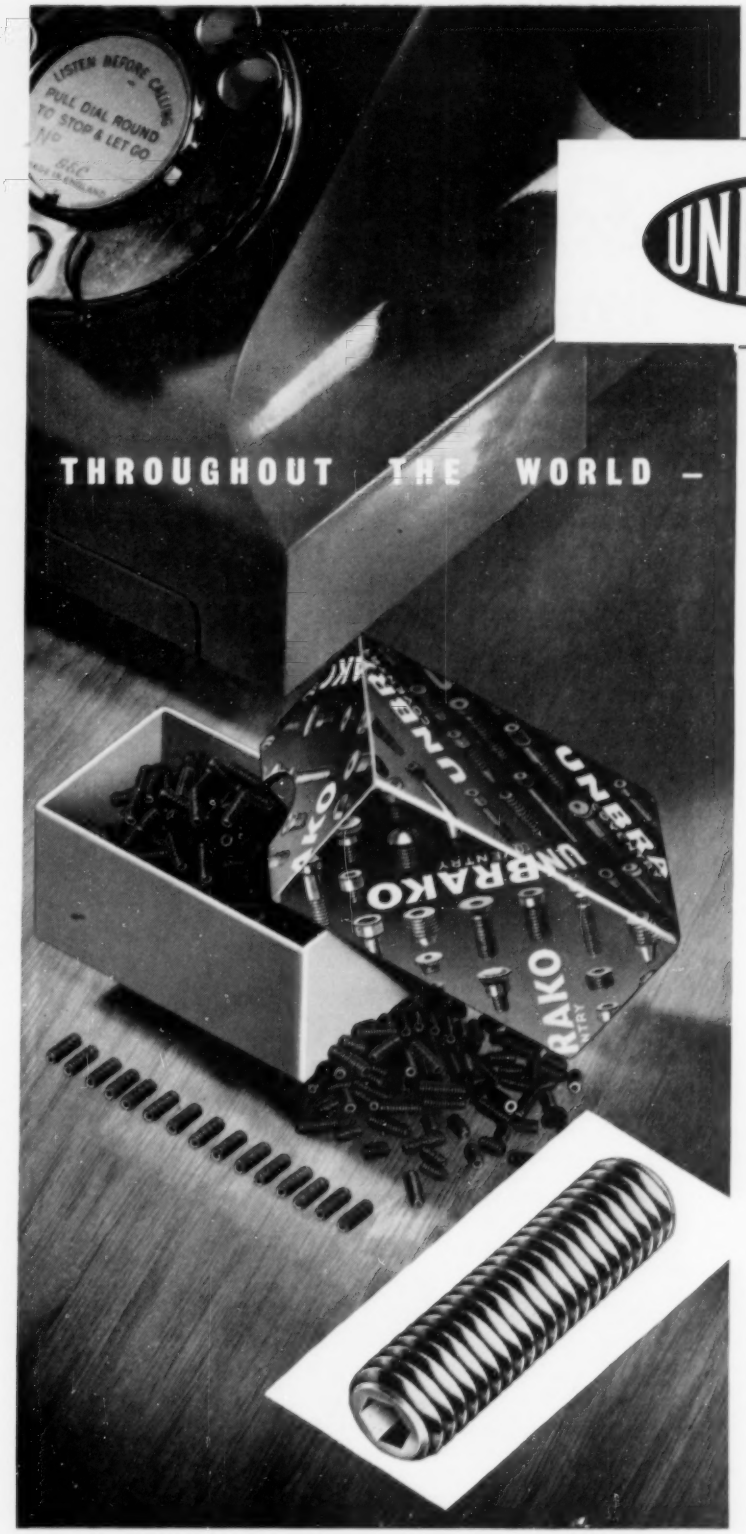
A SUBSIDIARY OF THE GENERAL ELECTRIC CO. LTD. OF ENGLAND



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G.E.C.

improved xenon rectifiers

G.E.C. Xenon rectifiers now readily available:

Type No.	Filt. Volt	Filt. Current	P.I.V	$i_{k(ave)}$	Equivalents	Comparable 'plug in' mercury rectifiers
GXU2	5.0V	7.1A	10kV	1.25A	4B32 RR3-1250 CV2518	
GXU3	4.0V	11.0A	13kV 10kV	1.25 A 1.5 A		CV5, GU20/21, AH221, ESU200, RG4—1250
GXU4	4.0V	7.0A	13kV	1.25 A		CV1629, RG3—1250

G.E.C. Xenon rectifiers incorporate these improvements:

1 'Boxed in' anode

2 Modified anode seal

3 Advanced cathode design

These give the following advantages:

a Longer average life

b Higher Peak inverse voltage

c Reduction of R.F. noise

d Less stringent smoothing requirements (with most-Xenon rectifiers reversing the phase of the anode supply affects the 'strike' voltage—typical variation is of the order of 4:1. If the circuit connections are such that the higher 'strike' voltage is operative then additional smoothing is necessary. G.E.C. Xenon rectifiers do not show this effect).

Full particulars of these Valves can be obtained from the G.E.C. Valve and Electronics Dept.

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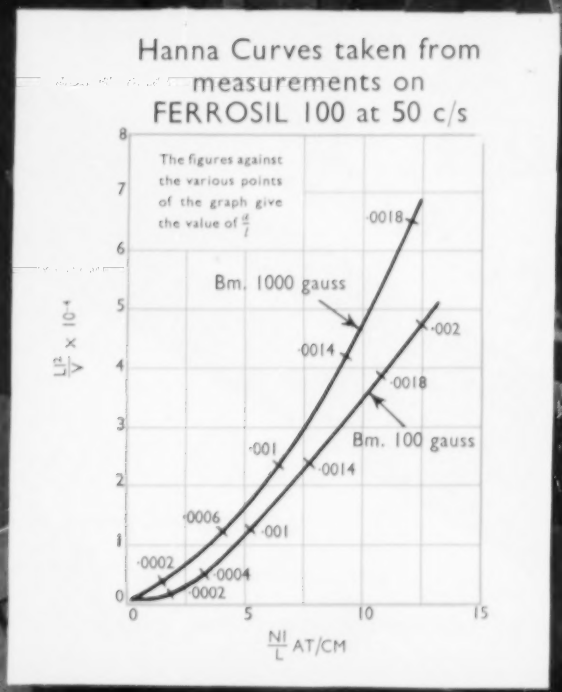
THE GENERAL ELECTRIC CO. LTD., MAGNET HOUSE, KINGSWAY, LONDON, W.C.2



'Hanna' curves for 'FERROSIL 100'

Typical data of 'Ferrosil 100' stampings is shown by Hanna Curves such as are widely used by electronic engineers in the design of iron cored chokes and transformers carrying both A.C. and D.C. currents. Low-loss high-permeability Ferrosil stampings are ideally suited to an infinite number of applications, of which only a few are shown. Not only is there an R.T.B. stamping for every job, but a wealth of technical data is available both in our new catalogue and by request from

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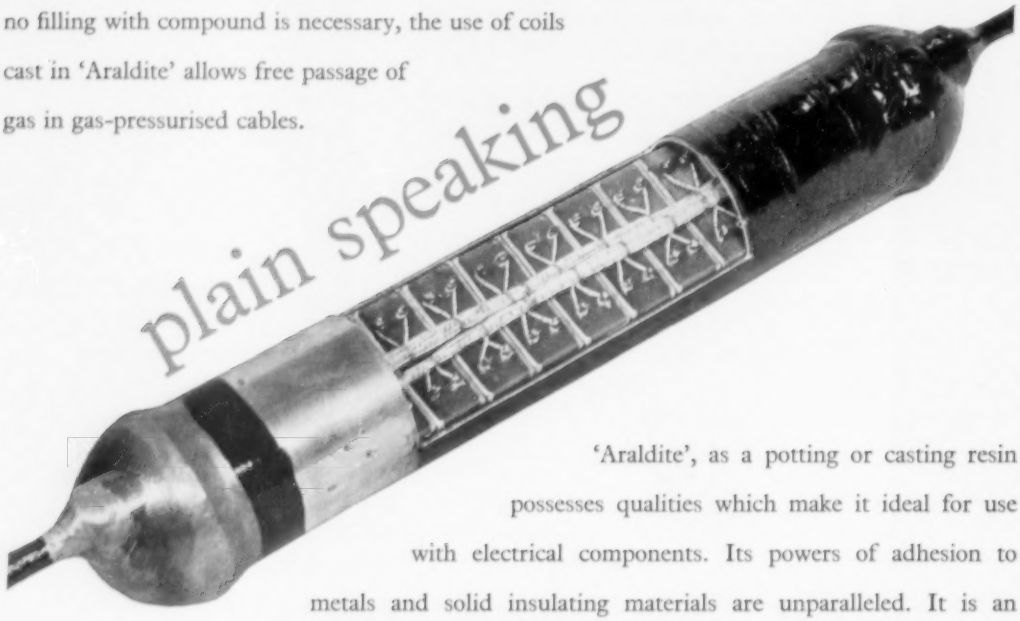
RTB

These Mullard Loading Coils represent the latest technique in splice loading of voice frequency cables, whereby the exact number of coils may be installed to meet current requirements. The coils are cast in 'Araldite' epoxy resin which gives complete protection against climatic conditions and permits the use of shapes which will occupy the minimum of space in the joint. As no filling with compound is necessary, the use of coils cast in 'Araldite' allows free passage of gas in gas-pressurised cables.



'Araldite', as a potting or casting resin possesses qualities which make it ideal for use with electrical components. Its powers of adhesion to metals and solid insulating materials are unparalleled. It is an excellent insulator, tough, flexible, highly resistant to 'tracking' and has a low power factor. It does not cause corrosion, provides complete protection against tropical conditions and complies with service requirements relating to sealing of components.

'Araldite' epoxy resins have a remarkable range of characteristics and uses.

They are used

- ★ **for bonding metals, porcelain, glass, etc.**
- ★ **for casting high grade solid insulation.**
- ★ **for impregnating, potting or sealing electrical windings and components.**
- ★ **for producing glass fibre laminates.**
- ★ **for producing patterns, models, jigs and tools.**
- ★ **as fillers for sheet metal work.**
- ★ **as protective coatings for metal, wood and ceramic surfaces.**

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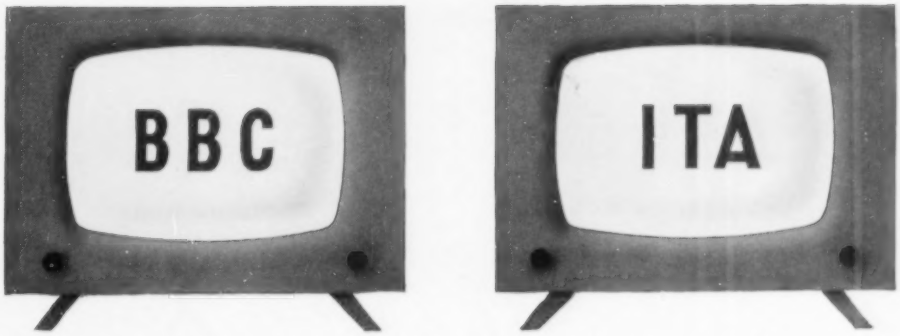
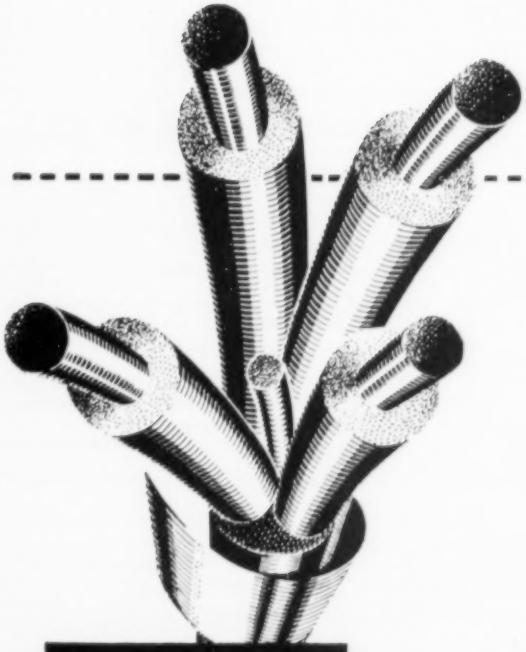


Two Programmes on one cable

With the minimum of complexity



Our manufacturing specification for television relay cables ensures very low cross-talk couplings within the cable. This provides for satisfactory common carrier operation with consequent simplification of system design. Further details of the BICC standard range of TR8 cables are contained in Publication TD T20 available on request.

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Polythene insulated and sheathed, screened quad cables.

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Small they may be, but Welwyn Insulated High Stability Carbon Resistors are passed out with the exceptionally high H1 rating. It means enduring 84 days under conditions of extreme exposure to tropical heat and humidity —but these Welwyn resistors can take it, without even the smallest change beyond the allowed deviation. *That their climatic performance puts them in the H1 class is a rare distinction for this kind of resistor, and a tribute to the high standards of 'quality control' that prevail at Welwyn Electrical Laboratories.*

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C81	... ¼ Watt	...	$\frac{11}{16}$" × $\frac{5}{16}$" diameter
C82	... ½ Watt	...	$\frac{7}{8}$" × $\frac{1}{4}$" diameter
C83	... 1 Watt	...	$1\frac{1}{16}$" × $\frac{3}{8}$" diameter

Insulation resistance exceeds $10^{12}\Omega$ at 500 volts DC

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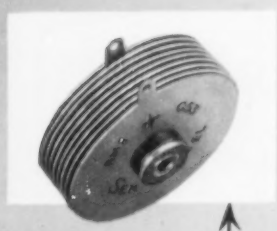

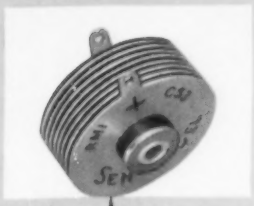

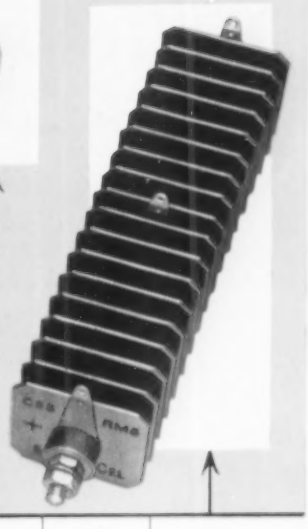

TYPE	RM0	RM1	RM2	RM3	RM4	*RM5
Maximum ambient temperature	35°C 55°C	35°C 55°C	35°C 55°C	35°C 55°C	40°C 55°C	40°C 55°C
Maximum output current (mean)	30mA 15mA	60mA 30mA	100mA 60mA	120mA 90mA	250mA 125mA	300mA 150mA
Maximum input voltage (r.m.s.)	125V	125V	125V	125V	250V	250V
Maximum peak inverse voltage	350V	350V	350V	350V	700V	700V
Max. instantaneous peak current	Unlimited	Unlimited	Unlimited	Unlimited	Unlimited	Unlimited
Weight	0.82 oz.	1 oz.	1.4 oz.	2 oz.	4.5 oz.	4.75 oz.

*For use in voltage doubler circuits the peak inverse and maximum input voltages are halved, current output being as for half wave operation.

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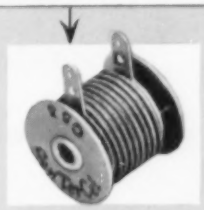

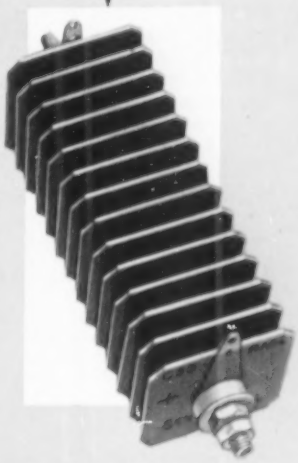

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Ratio: 2 : 1+1 (centre-tapped)
Primary Inductance: 10 H at 1.6 mA (1000 c.p.s.)
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COMPONENT DIVISION

FORTIPHONE LIMITED

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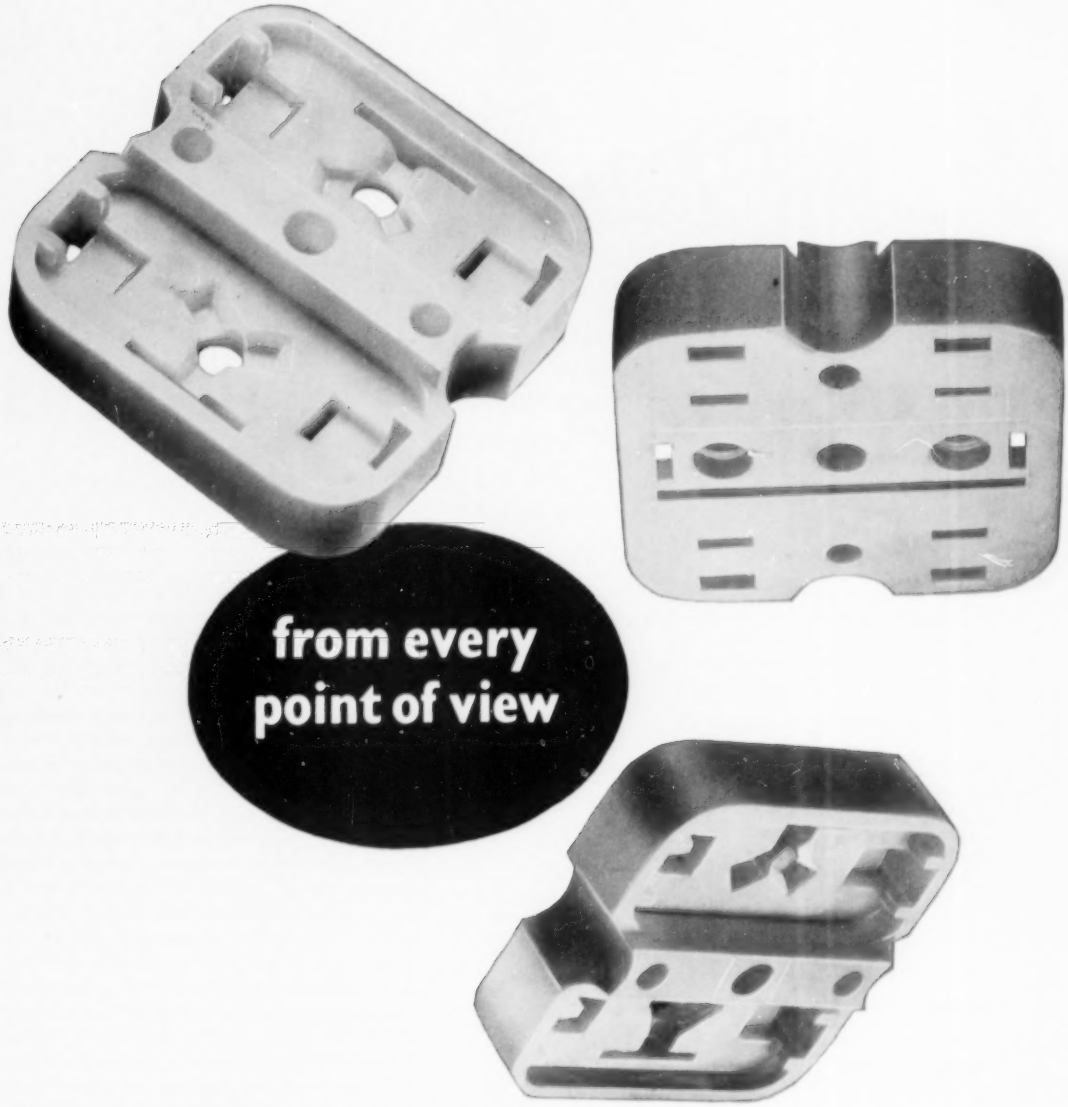

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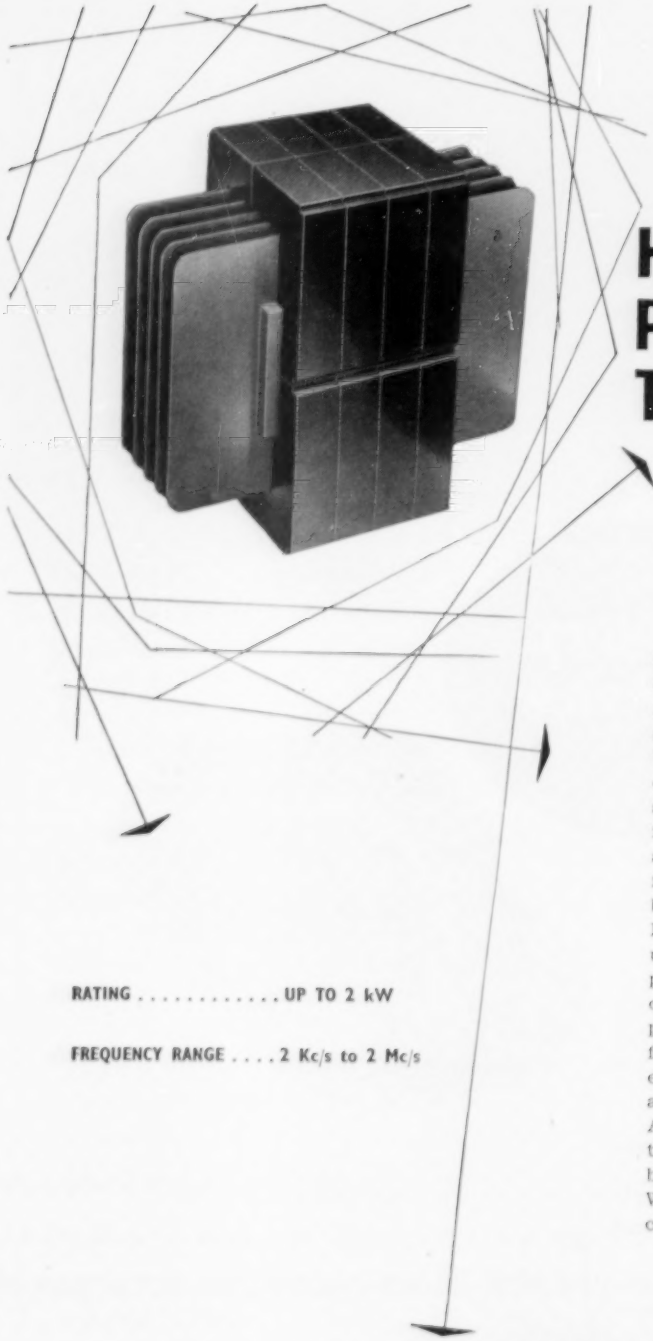

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WIRELESS ENGINEER

Vol. 33 AUGUST 1956 No. 8

DISC-SEAL TRIODE AMPLIFIERS

By G. Craven

(Standard Telecommunication Laboratories, Enfield)

SUMMARY.—The design of a resonant π-type coupling network for disc-seal triodes operating in the earthed-grid connection at frequencies in the range 300–3,000 Mc/s is considered. A coaxial form of line is adopted. Tuning for a small range can be by a "screw" or, for a larger range, by a built-in capacitance.

Complete amplifiers can give 100-W output and 30-dB gain using three stages.

Introduction

DISC-SEAL, or planar-element, triodes are commonly used as amplifiers in the frequency range 300–3,000 Mc/s and may be regarded as intermediate between conventional valves, employed at lower frequencies, and the velocity-modulation types found necessary at still higher frequencies. As is appropriate to this frequency range, the associated circuitry is provided by rigid coaxial lines, the valve being so constructed that, electrically, it can be made an integral part of the line. The common-grid (earthed-grid) type of circuit is generally employed, the input and output circuits being tuned to resonance by adjusting the length of the line so that, electrically, it is some odd multiple of a quarter wave. Radio-frequency power is introduced and abstracted by probes or loops that are connected to flexible coaxial lines. By providing feedback the amplifier may be readily converted to an oscillator. Examples of the circuitry and general construction are to be found in the literature[1,2].

Though capable of good performance, amplifiers of this type have a number of disadvantages, particularly where a multistage amplifier is contemplated. The number of adjustments per stage is excessive, the plunger contacts are troublesome and the complete amplifier is unnecessarily large. This is difficult to avoid if a large tuning range is needed. However, if only limited frequency coverage is required radical improvement is possible. The plunger tuning may be eliminated and replaced by some type of capacitive arrangement. Then, provided a suitable interstage network is employed, the multistage amplifier can be constructed as an integral unit and considerable simplification obtained.

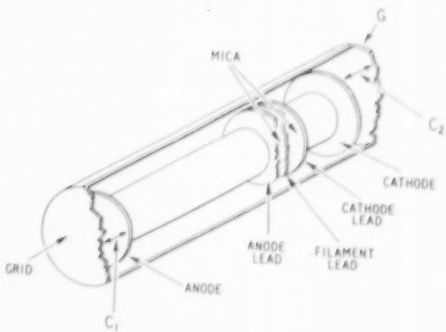

Fig. 1. *Basic construction of half-wave resonant line for disc-seal triode.*

A coupling network for disc-seal triodes should have the following properties:—a large impedance transformation with good efficiency (the output and input impedances of driver and driven valve may easily be 10,000 ohms and

MS accepted by the Editor, September 1955

50 ohms, respectively); wide bandwidth; simple construction; and the introduction of power-supply leads and forced-air cooling should not affect the circuit operation.

A network that fulfils these requirements in some degree is that based on a half-wave resonant line, as shown in Fig. 1. Though normally of unity transformation, capacitive end-loading modifies this greatly; it may, like its lumped-circuit analogue, the resonant π, approach an impedance ratio determined by the square of the capacitance ratio. The maximum bandwidth obtainable with this circuit depends on a number of things but chiefly the anode-grid capacitance and anode-load resistance. Practical features include the introduction of power-supply leads and, where necessary, forced-air cooling at the voltage node point. Where forced cooling is not required the inner conductor provides a mass of metal in contact with the anode to limit rapid temperature changes. Suitably sized holes may be drilled in the outer conductor to permit air circulation. Exact tuning to resonance may be made by a capacitive screw parallel to the electric field at the high impedance end. Where operation over a wider range of frequencies is required a larger variable capacitance, as shown in Fig. 2, can be used. This design has the advantage of eliminating moving contacts. The greater coverage is obtained at the expense of bandwidth.

Analysis

Useful design equations can be derived by solving the circuit equations for the network, redrawn in Fig. 3. The A matrices for these four-poles connected in tandem, assuming a lossless transmission line, are:

$$\begin{Vmatrix} V_1 \\ I_1 \end{Vmatrix} = \begin{Vmatrix} 1 & 0 \\ jB_1 & 1 \end{Vmatrix} X \begin{Vmatrix} \cos\theta & jZ_0 \sin\theta \\ j(\sin\theta)/Z_0 & \cos\theta \end{Vmatrix} X \begin{Vmatrix} 1 & R \\ jB_2 & jRB_2 + 1 \end{Vmatrix} X \begin{Vmatrix} V_2 \\ I_2 \end{Vmatrix} \quad .. \quad .. \quad (1)$$

where θ = electrical length of the transmission line: $= \dfrac{2\pi L}{\lambda}$;

L = physical length

λ = free-space wavelength

$B_1 = \omega C_1$; $B_2 = \omega C_2$; Z_0 = characteristic impedance of line as drawn in Fig. 3.

R = valve input resistance, $V_2 = 0$

Multiplying these matrices, beginning from the right:

$$\begin{Vmatrix} V_1 \\ I_1 \end{Vmatrix} = \begin{Vmatrix} 1 & 0 \\ jB_1 & 1 \end{Vmatrix} X \begin{Vmatrix} (\cos\theta - Z_0B_2\sin\theta)[jZ_0\sin\theta + R(\cos\theta - Z_0B_2\sin\theta)] \\ j[(\sin\theta)/Z_0 + B_2\cos\theta][jR(\sin\theta)/Z_0 + \cos\theta(1 + jRB_2)] \end{Vmatrix} X \begin{Vmatrix} V_2 \\ I_2 \end{Vmatrix} \quad (2)$$

Completing the multiplication and taking the A_{22} term only we have:

$$I_1/I_2 = \cos\theta - Z_0B_1\sin\theta + jR[\cos\theta(B_1 + B_2) + \sin\theta\,(1/Z_0 - Z_0B_1B_2)] \quad .. \quad .. \quad (3)$$

The conditions for resonance may be obtained by equating $-B_1$ to the susceptance seen to the right of it in Fig. 3. The admittance, Y, to the right of B_1 is given by A_{22}/A_{21} in the right-hand matrix of (2); i.e.,

$$Y = \frac{jR\left\{(\sin\theta)/Z_0 + B_2\cos\theta\right\} + \cos\theta}{jZ_0\sin\theta + R(\cos\theta - Z_0B_2\sin\theta)} \quad .. \quad (4)$$

Rationalizing, and taking the imaginary part only, for resonance,

$$-B_1 = \frac{(\sin\theta + Z_0B_2\cos\theta)(\cos\theta - Z_0B_2\sin\theta)/Z_0 - Z_0(\sin\theta\cos\theta)/R^2}{(\cos\theta - Z_0B_2\sin\theta)^2 + Z_0^2(\sin^2\theta)/R^2} \quad .. \quad .. \quad (5)$$

Substituting for B_1 in (3) and simplifying

$$I_1/I_2 = \frac{\cos\theta - Z_0B_2\sin\theta + jZ_0(\sin\theta)/R}{(\cos\theta - Z_0B_2\sin\theta)^2 + Z_0^2(\sin^2\theta)/R^2} \quad .. \quad .. \quad (6)$$

or calling T, the current transformation I_2/I_1 at resonance we have

$$|I_2/I_1|^2 = T^2 = (\cos\theta - Z_0B_2\sin\theta)^2 + Z_0^2(\sin^2\theta)/R^2 \quad .. \quad .. \quad (7)$$

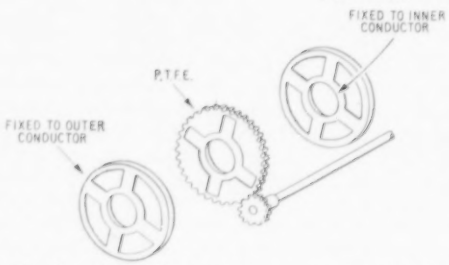

Fig. 2. *A variable capacitance in coaxial circuit.*

Design of Network

The design of a network requires equations (5) and (7). Generally, the practical problem is to obtain a specified transformation at a given frequency, in which case the equations must be rearranged in terms of whatever variables are chosen. Though four are available—Z_0, θ, B_1 and B_2—only three are really suitable, since

large variations in the characteristic impedance, Z_0, are not practicable. The susceptances, B_1 and B_2, in conjunction offer the greatest design flexibility, since for a given transformation, the electrical length, θ, may then be any value less than a certain maximum, the shorter lengths being obtained with decreased bandwidth. The second and third cases occur when θ is employed in conjunction with either B_1 or B_2.

In obtaining simple design equations for the various cases it is helpful to make the approximation that both the resonant frequency, f, and transformation, T, are unaffected by the load resistance, R. Generally this will lead to only small error because, in the range of frequencies considered, the reactance of valve inter-electrode capacitances is such as to make the circuit a resonant device of reasonably high loaded Q. Where desired, the values obtained from the approximate equations may be substituted in the exact equations and better approximations obtained. However, this should not generally be necessary. Discarding the terms containing R in (5) and (7) we have:

$$-B_1 = \frac{\sin\theta + Z_0B_2\cos\theta}{Z_0(\cos\theta - Z_0B_2\sin\theta)} \quad .. \quad (5a)$$

and $$T = \cos\theta - Z_0B_2\sin\theta \quad .. \quad .. \quad (7a)$$

or $$-B_2 = \frac{T - \cos\theta}{Z_0\sin\theta}$$

Since $B_1 = \omega C_1$ and $B_2 = \omega C_2$ the design capacitances are:

$$-C_2 = \frac{T - \cos\theta}{\omega Z_0\sin\theta} \quad .. \quad .. \quad .. \quad (8)$$

$$-C_1 = \frac{\sin\theta + Z_0\omega C_2\cos\theta}{\omega Z_0 T} \quad .. \quad (9)$$

In designing the network an experimental value of θ is chosen and the current transformation, T, decided on (its sign will be negative). Equ. (8) is then solved and the result substituted in (9). If too large a value of θ is chosen, capacitance values that are zero or negative will be obtained. This method has its main application at the lower frequencies when it is desired to restrict the length of an amplifier by adding capacitances in shunt with the valve capacitances.

The case where the length, θ, is a parameter in conjunction with either C_1 or C_2 remains to be considered. The choice between C_1 and C_2 as the other parameter depends on practical considerations: the current transformation, T, required, the anode–grid capacitance of the driver valve, and the cathode–grid capacitance of the driven valve. Usually, the input capacitance of the latter will not be high enough to obtain the desired transformation, even when the driver valve output capacitance is quite small. In this case C_2 must be considered as the parameter to be employed with θ. Less frequently, if the required transformation is small and the input capacitance of the driven valve high, C_1 must be considered as the variable to be used with θ. Taking Equs. (5a) and (7a) and eliminating Z_0B_2 from their sum,

$$\cos\theta - Z_0B_1\sin\theta = 1/T$$

or $$\theta = \cos^{-1}\{(\cos\phi)/T\} - \phi \quad .. \quad .. \quad (10)$$

where $\phi = \tan^{-1} Z_0B_1$

Similarly, when θ and C_1 are the parameters:

$$\theta = \cos^{-1}(T\cos\psi) - \psi \quad .. \quad .. \quad (11)$$

where $\psi = \tan^{-1} Z_0B_2$

Equs. (10) and (11) give the maximum line length, in each case, which can provide the desired transformation and, since the minimum necessary capacitance is added, the widest bandwidth obtainable with this network. The value of C_2 or C_1 is of course obtained as before in Equs. (8) and (9), respectively.

One additional piece of information is required for constructional purposes: the position of the voltage node. Power supply leads may be introduced at this point which, since C_1 and C_2 are presumed unequal, is displaced from the centre. Calling the line lengths (to C_1 and C_2) on either side of the node θ_1 and θ_2, each combination θ_1C_1, θ_2C_2, as seen from the nodal point, must be series resonant; i.e. the susceptance is,

$$\tan(\theta_1 + \tan^{-1} Z_0B_1)/Z_0 = \infty \quad .. \quad (12)$$

$$\tan(\theta_2 + \tan^{-1} Z_0B_2)/Z_0 = \infty \quad .. \quad (13)$$

The distance at which the node occurs from the particular terminating capacitor is then:

$$\theta_1 = \pi/2 - \tan^{-1} Z_0B_1 \quad .. \quad .. \quad (12a)$$

$$\theta_2 = \pi/2 - \tan^{-1} Z_0B_2 \quad .. \quad .. \quad (13a)$$

Either (12a) or (13a) may be used.

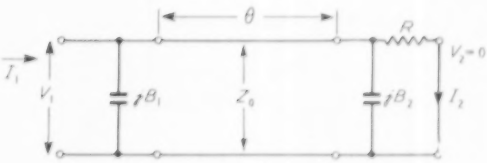

Fig. 3. *Equivalent circuit of half-wave line and valve.*

Example

A valve with an effective anode–grid capacitance of 3 pF, and requiring an anode load of 2,500 ohms, is to drive a valve with an input resistance of 100 ohms, and an effective cathode–grid capacitance of 10 pF. The design frequency is 300 Mc/s. There are no severe restrictions on length.

First, a suitable characteristic impedance must be chosen. Low values of Z_0 require a large value of θ, without any compensating advantage, while a high value of Z_0 results in large outer conductors. Choosing an intermediate value, let $Z_0 = 50\ \Omega$.

The transformation, T, required is,

$$T = -\sqrt{\frac{2500}{100}} = -5.$$

Since the amplifier may be of any reasonable length, θ may be used as a parameter in conjunction with a capacitance at either end.

Let $C_1 = 3{\cdot}5$ pF (this permits a small variable capacitance to obtain exact tuning).

θ and C_2 are to be found.

Substituting for ϕ in (10)

$$\theta = \cos^{-1}\{-(\cos 18{\cdot}25^\circ)/5\} - 18.25^\circ$$

$$\theta = 82{\cdot}7^\circ \text{ or } L = \theta\lambda/2\pi = 23 \text{ cm}.$$

Substituting for θ in (8)

$$C_2 = 54{\cdot}9 \text{ pF}$$

Since the cathode–grid capacitance is 10 pF an additional 44·9 pF is required in shunt with it. Capacitances of this type are conveniently constructed from a section of very low characteristic impedance line, the equivalent capacitance, C, being given, for short lengths by:

$$C = \frac{100L_c}{3Z_{03}} \qquad \ldots\ \ldots\ \ldots\ \ldots \qquad (14)$$

where C is in pF, Z_{03} is the characteristic impedance and L_c the length of the section in cm. Where necessary, correction for the discontinuity susceptance[3] may be made.

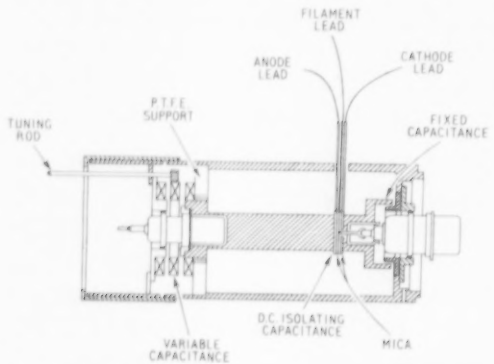

Fig. 4. *Detailed construction of line.*

Continuing with the example, the voltage node occurs at a distance θ_2 from C_2:

$$\theta_2 = \pi/2 - \tan^{-1} Z_0\omega C_2 = 10{\cdot}9^\circ$$

or 3·0 cm from C_2.

The error resulting from neglecting the resistance terms in (5) and (7) may now be checked. Substituting for θ and C_2 we have the corrected values: $T = -5{\cdot}03$ and $C_1 = 3{\cdot}53$ pF, which is a negligible error.

The constructional details of a typical amplifier are shown in Fig. 4, which is an example of a design where the length has been reduced by increasing C_1 and C_2 beyond the minimum required to obtain the desired transformation. The variable capacitance, which contributes the major part of C_1, is of the type shown in Fig. 2. The inner conductor is supported at the driver valve end by a polytetrafluorethylene (P.T.F.E.) ring and at the other end by a fixed capacitance (the major part of C_2), which is free to move small amounts longitudinally, in order to take up expansion effects. Generally speaking, polythene insulation is to be avoided at the anode end of the line on account of the high operating temperatures. Power-supply leads are taken out at the zero potential point and, if the capacitance shown is of adequate value, additional by-passing external to the unit is not needed. If fixed bias is preferred to cathode bias, the usual type of annular grid capacitance may be constructed. However, this represents a common impedance between input and output circuits which can sometimes be troublesome. An additional point evident from Fig. 4 is that the external fixed capacitance at the cathode is not in shunt with the valve cathode–grid capacitance, but separated from it by a short section of line. In most cases this approximation is unlikely to lead to serious error. However it may be necessary to take it into account at the higher frequencies. Under these circumstances the input resistance and capacitance employed in the design equations are the equivalent values seen at the plane of the cathode terminal of the valve.

Conclusions

The network should be useful throughout the range of frequencies in which disc-seal triodes are applicable: about 300–3 000 Mc/s.

The transformation ratio tends, for large values of C_0, to $T = -\frac{C_2}{C_1}$; hence almost any desired transformation can be obtained. Dissipative loss is the practical limitation, but as this is quite small in coaxial line networks it should not constitute a serious restriction.

Throughout the paper reference to disc-seal triodes has been made exclusively. However, where suitable tetrodes are available the application is obvious. Generally, the transformation required should be less.

The type of construction shown in Fig. 4

readily lends itself to multistage amplifiers in which any number of stages can be built as a relatively compact integral unit. For example a three-stage amplifier of 30 dB gain, 100-watt output, operating at the low-frequency end of the 300–3,000 Mc/s range was approximately 13 in. long and 5 in. diameter.

The bandwidth obtainable can be found from equations derived in the appendix. Experience indicates that it is sufficient for a great many purposes. Where larger bandwidths than can be obtained with this network are required a coaxial construction based on Shea's impedance transforming filter is to be preferred[4].

Acknowledgments

The writer wishes to thank Electrical and Musical Industries for permission to publish this paper; also Mr. K. D. Mills, now of Canadian Westinghouse, Hamilton, Ont., for many helpful discussions on disc-seal triode amplifiers.

APPENDIX

Bandwidth

The frequency band over which the transformation is substantially constant can be calculated with the aid of Equ. (3). As discussed in the text the imaginary part of this equation is practically zero at resonance and so, if we define the cut-off frequencies as those at which the real and imaginary parts are equal, we have:

$$R\left[B_1\left(\frac{\cos\theta + (\sin\theta)/Z_0B_1}{\cos\theta - Z_0B_1\sin\theta}\right) + B_2\right] = \pm 1 \quad .. \quad (15)$$

or

$$R[\tan(\theta+\phi)/Z_0 + B_2] \pm 1 = 0 \quad .. \quad .. \quad (16)$$

where the notation is the same as that employed in (1) and (10).

Since the angular frequency, ω, is contained in θ, ϕ and B_2 the solution is rather tedious and an approximation giving the bandwidth directly is useful.

Putting $\omega = \omega_0(1+\delta)$, where ω_0 = angular resonant frequency, and substituting in (16)

$$R\left[\frac{\tan\left\{L\omega_0(1+\delta)/c + \tan^{-1}Z_0C_1\omega_0(1+\delta)\right\}}{Z_0} + \omega_0C_2(1+\delta)\right] \pm 1 \approx 0 \quad .. \quad .. \quad (17)$$

where L and c are the line length and velocity of light, respectively.

Assuming δ small ($\delta\tan\theta \approx \delta\theta$) and $\omega_0C_1Z_0$ a small angle, we have, expanding and neglecting δ^2 terms

$$R\left(\frac{1}{Z_0}\left[\begin{array}{l}\tan\left(\frac{\omega_0L}{c}+\tan^{-1}Z_0\omega_0C_1\right)+\delta\omega_0\left(\frac{L}{c}+Z_0C_1\right)\\ +\delta\omega_0\left(\frac{L}{c}+Z_0C_1\right)\tan^2\left(\frac{\omega_0L}{c}+\tan^{-1}Z_0\omega_0C_1\right)\\ +\omega_0C_2+\delta\omega_0C_2\end{array}\right]\right) \pm 1 \approx 0 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (18)$$

And since $\left\{\tan(\omega_0L/c + \tan^{-1}Z_0\omega_0C_1)\right\}/Z_0 + \omega_0C_2 = 0$ and the response about resonance, to the order considered, is symmetrical we have:

$$2\delta\omega_0/2\pi = 2\delta f = \text{bandwidth} \approx \frac{1}{\pi R\left[\left\{L/cZ_0+C_1\right\}\left\{1+\tan^2(\omega_0L/c+\tan^{-1}Z_0\omega_0C_1)\right\}+C_2\right]} \quad .. \quad .. \quad (19)$$

Substituting the values from the example in the text in this equation gives: bandwidth = 5·5 Mc/s which is less than 10 per cent in error compared with the result obtained by trial solution of (16).

The main factors affecting bandwidth are evident from Equs. (16) and (19). The term in the square brackets of (16) is zero at resonance; hence for large values of B_2, $\tan(\theta+\phi)$ is also large and this adversely affects bandwidth, (19). Obviously a narrow bandwidth results from a large transformation and both high anode load resistance and capacitance [these all lead to large values of $\tan(\theta+\phi)$]. Similarly, (16) shows that high values of Z_0 increase bandwidth so that where the latter is important the highest characteristic impedance consistent with reasonable dimensions should be chosen.

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4. Pat. No. 588,292.

KLYSTRON CONTROL SYSTEM

By R. J. D. Reeves

(E. K. Cole, Ltd., Malmesbury, Wilts)

(Concluded from p. 167, July issue)

Part 3 (contd.)—Specimen Design for A.F.C. Frequency-Control Loop as a Sampling Servo

THE behaviour of the frequency-control loop as a sampling servo may be investigated without considering the influence of the mode-centring loop, to which it is implicitly coupled through the klystron. This is permissible because we can confine the interpretation of the results to that part of the error-frequency spectrum which is outside the province of the mechanical control or, alternatively, we can postulate that the magnitude of the reflector-voltage deviations that occur produce insufficient mode-centring errors to energize the centre-stable polarized relay in the motor circuit.

In this analysis the method of the z transform will be used whereby the substitution

$$z = \epsilon^{p\tau} \quad .. \quad .. \quad .. \quad .. \quad (12)$$

is employed, τ being the repetition period.

From Equ. (9)

$$E_b = -\frac{1}{kC}\frac{\alpha}{p(p+\alpha)}(I_A - I_B) \quad .. \quad (13)$$

where $\alpha = 1/CR$

From the z transform tables of Barker[2] the pulse transfer function (p.t.f.) of this control circuit is:

$$W(z) = -\frac{m\tau}{kC}\cdot\frac{(1-d)z}{(z-1)(z-d)} \quad .. \quad (14)$$

where $d = \epsilon^{-\tau/CR}$

and the duty ratio m is introduced so that I_A and I_B can be given their direct-current values.

Now if $I_A - I_B$ is considered to be related to the Miller output voltage increments by a constant transfer conductance β then the complete loop p.t.f. is:

$$W_l(z) = -\frac{m\beta\tau}{kC}\cdot\frac{(1-d)z}{(z-1)(z-d)} \quad .. \quad (15)$$

and the external p.t.f. of the feedback system is:

$$W_f(z) = \frac{W(z)}{1 - W_l(z)} = -\frac{\frac{m\tau}{kC}(1-d)z}{z^2 - (1+d)z + d + \frac{m\beta\tau}{kC}(1-d)z} \quad .. \quad .. \quad (16)$$

The characteristic function $Q(z)$ is:

$$z^2 + \left\{\frac{m\beta\tau}{kC}(1-d) - (1+d)\right\}z + d \quad (17)$$

Since the substitution (12) transforms the left-half p-plane to the interior of unit circle in the z-plane, the stability criterion requires all roots in the z plane to have a modulus less than unity.

This is succinctly expressed by the triad:

$$Q(1) > 0,\ Q(-1) > 0,\ Q(0) < 1 \quad .. \quad (18)$$

Only the second of these is non-trivial here. In this,

$$1 + (1+d) + \frac{m\beta\tau}{kC}(d-1) + d \text{ to be} > 0$$

or
$$2(1+d) + \frac{m\beta\tau}{kC}(d-1) > 0$$

or
$$\frac{m\beta\tau}{kC} < 2\frac{1+d}{1-d}$$

i.e.,
$$\beta < \frac{2kC}{m\tau}\cdot\frac{1+\epsilon^{-\tau/CR}}{1-\epsilon^{-\tau/CR}} \quad .. \quad .. \quad (19)$$

Substituting the values from Fig. 13 we have:

$$\tau/CR = 1; \qquad kC = 0{\cdot}002\ \mu\text{F};$$
$$\tau = 500\ \mu\text{sec}; \qquad m = 0{\cdot}001;$$

therefore, for stability, β must be less than 17 mA/V. The factor β is the product of a number of transfer functions many of which cannot be held particularly constant, so that it is desirable to work with a nominal value very much less than that stability limit. If the loop gain variation can be assessed then a maximum safe value for β can be computed.

It should be noted that in a comprehensive a.f.c. system of this type, it can be assumed that the reflector is working at mode centre, and the modulation coefficient in Mc/s per volt is not subject to variations due to permanent deviations from the centre. This removes one source of variations in β.

Further insight into the behaviour of the control loop may be obtained by finding the largest value of β which may be employed without over-correction; i.e., the critically-damped condition. The criterion for this is that the characteristic equation shall exhibit a double root on the real p-axis, and as the transformation (12) maps the real p-axis on to the positive half of the real z-axis, the characteristic equation in z [$Q(z) = 0$] will exhibit a corresponding double root.

$Q(z)$ is quadratic in z and has equal real roots when:

$$\left\{\frac{m\beta\tau}{kC}(1-d) - (1+d)\right\}^2 = 4d$$

i.e.,

$$\left(\frac{m\beta\tau}{kC}\right)^2 - 2\left(\frac{1+d}{1-d}\right)\frac{m\beta\tau}{kC} + 1 = 0 \qquad (20)$$

This is quadratic in β and, when the previously assumed values of τ and kC are substituted, the solutions are:

$\beta = 16{\cdot}3$ or $1{\cdot}0$

The first of these is near the stability boundary previously determined by equation (19) but it is the lower value of β which corresponds to a positive real root of $Q(z)$ and signifies critical damping.

The onset of ringing here has little to do with the sampling process but is a consequence of the fact that the loop is dominated by an integrator and one time constant. A similar loop with an attenuator m substituted for the sampling switch would also start to ring at approximately this value of β. Clearly, the extraneous time constant is not desirable if it is necessary to work near the shortest possible correction time compatible with the repetition time but, in this case, the repetition frequency is rather high (2 kc/s) and in the design of the system the correction rate was not a point at issue. Moreover, loop ringing from this cause is not an indication of approaching instability nor, in moderation, is it an undesirable property in an a.f.c. loop; so it is by no means essential to restrict β to values less than this. Taking $\beta = 1$ as a nominal design centre, the external pulse transfer function simplifies to:

$$W_f(z) = \frac{m\tau}{kC}\,\frac{(1-d)z}{(z-\sqrt{d})^2} \qquad (21)$$

and this responds to unit impulse and unit step in the manner shown in Fig. 24.

If the time constant were eliminated, for example by interposing a cathode follower between the difference circuit and the Miller integrator, then it would be possible to improve the response time by a factor of two before approaching sampling limitations.

Mode-Centre Discriminator Characteristic

Unlike the frequency discriminator, the factors contributing to the pull-in characteristic of the mode-centring system are to a considerable extent outside design control. The mode shape has to be accepted per se and the sinusoid amplitude which, together with the mode width determines the capture range, needs to be as large as possible with due regard to the risk of pulling on to an adjacent mode during warm-up. A suitable amplitude for use with the R5222 klystron has been found to be about two mode widths (approximately 80 V) peak-to-peak. By assuming a parabolic form for the mode shape, an expression can be derived for the discriminator characteristic which should be reasonably descriptive of that actually realized, provided the crystal-current amplifier is a linear one and provided adjacent modes are spaced by at least two mode widths and do not oppose the pull in.

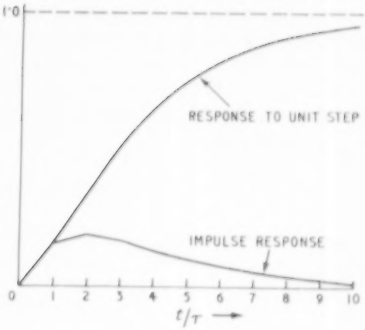


Fig. 24. *Computed transient response of frequency control loop.*

As described in Part 2, the mechanism of the discriminator is to project the mode shape on to a time scale in order that it may be operated on by a time discriminator. For the purpose of analysis, however, this process may be inverted and the time gates projected on to the voltage scale so as to eliminate the time variable. The relation between time and voltage is the sinusoid:

$$V = -V_s \sin \omega t \qquad 0 < t < 2\pi/\omega \qquad (22)$$

and it is clear that the early and late gates represent adjoining voltage intervals of length V_s with confluence at the mean reflector potential. But the integrating process in the time discriminator is carried out with respect to time, which is not linearly related to the reflector potential scale. The transformation is characterized by a weighting function derived from (22)

Thus $dV = \omega V_s \cos \omega t\, dt$

or

$$dt = \frac{dV}{\omega V_s \sqrt{1 - V/V_s}} \qquad (23)$$

The ω and V_s are scale factors; it is the term under the radical which influences the shape of the characteristic. As in the case of a pulsed frequency discriminator, the overall discrimination characteristic is not that of the discriminator itself but the product of the discriminator and the input 'spectrum', and the weighting function just derived acting on a parabolic mode produces a discrimination characteristic shown as A in Fig. 25.

To derive this, let V_m be the mode span and also the amplitude of the sinusoid; i.e., $V_s = V_m$. The effective discriminator law $f_1(V)$ is given by:

$$f_1(V) = \begin{cases} 0 & V < -V_m \\ (1 - V^2/V_m^2)^{-\frac{1}{2}} & -V_m < V < 0 \\ -(1 - V^2/V_m^2)^{-\frac{1}{2}} & 0 < V < V_m \\ 0 & V_m < V \end{cases} \qquad (24)$$

The mode shape is:

$$f_2(V) = \begin{matrix} 0 & V < -V_m/2 \\ 1 - 4V^2/V_m^2 & -V_m/2 < V < V_m/2 \\ 0 & V_m/2 < V \end{matrix} \quad (25)$$

The voltage scale can be normalized by putting $V/V_m = x$, and a misalignment variable μ introduced such that

$$\mu = \Delta/V_m \quad .. \quad .. \quad .. \quad (26)$$

where Δ is the voltage displacement between the reflector and mode centre.

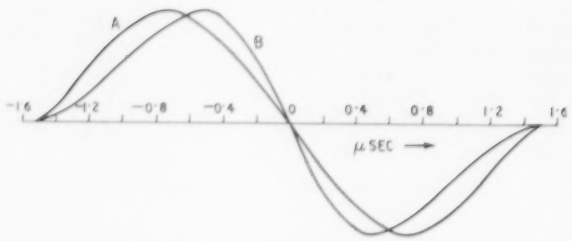


Fig. 25. *Computed characteristic of mode centre discriminator.*

Reference will now be made to Fig. 26.

$$F_1(x) = \frac{1}{\sqrt{(1-x^2)}} \text{ or } 0 \quad .. \quad .. \quad (27)$$

$$F_2(x) = \begin{matrix} 0 & x < \mu - \frac{1}{2} \\ 1 - 4(x-\mu)^2 & \mu - \frac{1}{2} < x < \mu + \frac{1}{2} \\ 0 & \mu + \frac{1}{2} < x \end{matrix} \quad (28)$$

The integrated product of these two curves is a function of μ and can be obtained in two parts, depending on whether μ is greater or less than $\frac{1}{2}$

Centre:

$$\int_{\mu-\frac{1}{2}}^{0} \frac{1-4(x-\mu)^2}{\sqrt{1-x^2}}\,dx - \int_{0}^{\mu+\frac{1}{2}} \frac{1-4(x-\mu)^2}{\sqrt{1-x^2}}\,dx \quad (29)$$

Valid for $0 < \mu < \frac{1}{2}$

Skirts:

$$\int_{\mu-\frac{1}{2}}^{1} \frac{1-4(x-\mu)^2}{\sqrt{1-x^2}}\,dx \quad (30)$$

Valid for $\frac{1}{2} < \mu < 1\frac{1}{2}$

The indefinite integral concerned is found to be

$$-(1+4\mu^2)\sin^{-1}x + 2(x-4\mu)\sqrt{1-x^2} \quad (31)$$

and substituting limits, the centre portion of the discrimination characteristic is given by:

$$\epsilon = -16\mu + (1+4\mu^2)\left[\sin^{-1}(\mu-\tfrac{1}{2}) + \sin^{-1}(\mu+\tfrac{1}{2})\right] + (6\mu-1)\sqrt{\tfrac{3}{4}-\mu-\mu^2} + (6\mu+1)\sqrt{\tfrac{3}{4}+\mu-\mu^2} \quad .. \quad (32)$$

and the skirts by:

$$\epsilon = (1+4\mu^2)\cos^{-1}(\mu-\tfrac{1}{2}) - (6\mu+1)\sqrt{\tfrac{3}{4}+\mu-\mu^2} \quad .. \quad .. \quad .. \quad (33)$$

This is the expression that produces the curve A of Fig. 25.

For comparison, the characteristic curve B that would result from an unweighted voltage discriminator (obtained from a rectilinear mapping waveform) is also drawn. It is seen that the sinusoidal discriminator exhibits a lower error sensitivity at the origin but, due to the weighted extremities, it produces a stronger pull-in signal. Since the main purpose of the large amplitude sinusoid is to accomplish a good pull-in performance this is considered a useful property.

By-Pass Gearbox and Flash-Back Mechanism

Wide-range frequency search is accomplished by the sawtooth motion of the klystron-cavity tuning plunger. When the required frequency is found, the tuning mechanism then forms part of a closed loop and a slower rate of tuning is required to ensure loop stability. To this end a by-pass gearbox has been developed which is effectively equivalent to the slipping-ball epicyclic type of tuning mechanism, and permits two distinct transmission ratios. Fig. 27 illustrates the method employed.

The input pinion (1) is secured to its shaft and transmits the drive through the spur reduction (2), (3), (4) to the output pinion (5). This provides the slow rate of tuning. Also on the shaft is a stack of tab washers, each of which is capable of a certain amount of lost motion before engaging the next one.

The input pinion engages the first of these washers, and sustained rotation in one direction will accumulate all of them until finally torque is transmitted to the ball clutch device (6). This clutch links the direct transmission to the

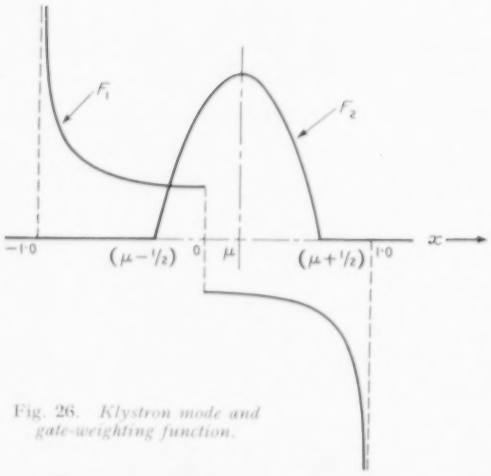


Fig. 26. *Klystron mode and gate-weighting function.*

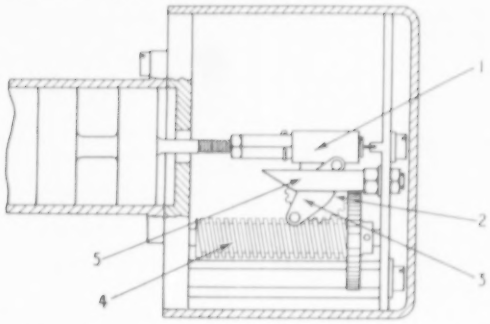
output pinion through two steel balls trapped in conical depressions in the clutch plates. The function of these is to resolve part of the applied torque into an axial thrust in order to push the output pinion out of engagement with gear (4) for the direct drive condition. A reversal of the motor removes this thrust and a return spring (not shown) then causes the output pinion to re-engage gear (4).

The features of this design to notice are that both the spur reduction and the low gear interval (number of tab washers) can be chosen freely, and at no time is excessive friction present in the mechanism. It can therefore be incorporated in small position-control systems without absorbing appreciable power from the motor.

A control system incorporating this type of mechanism exhibits the behaviour characteristic of a manual operation in that a rapid search results in an overshoot followed by a slow and careful approach to the correct setting from the other direction.

This mechanism would be an encumbrance in the search condition if the motor were called upon to reverse at the extremes of the piston travel, for then the low gear interval would have to be traversed before the return sweep could get properly under way. Another consideration which favours unidirectional search is that unless the motor were reversed by reversing the search bias in the control unit, there would be a phase ambiguity when the loop is closed. Consequently a flash-back mechanism has been adopted which simulates the performance of a step cam (with linear stroke), and returns the piston to the start of the sweep without interrupting the motor supply.

The functioning of this device is shown in Fig. 28. The piston is spring loaded against a travelling carriage (1) bearing a pinion (2) and crank (3) coupled together. The pinion engages with the worm (4) so that, as the carriage is pressed back by the piston spring, the crank rotates until one of the two pins it carries engages the peg (5). Rotation of the worm shaft then causes the carriage to travel with the crank pin sliding along the peg and, if the direction of travel is to the left in the figure, the pin approaches the extremity. Beyond this point, the pinion is free to turn and there is nothing to prevent the carriage being returned by the spring until the worm wheel revolves half circle and the other crank pin engages. The pinion diameter therefore determines the stroke. It can be seen that the flash back is a trigger action which does not depend on the motor once it is initiated.



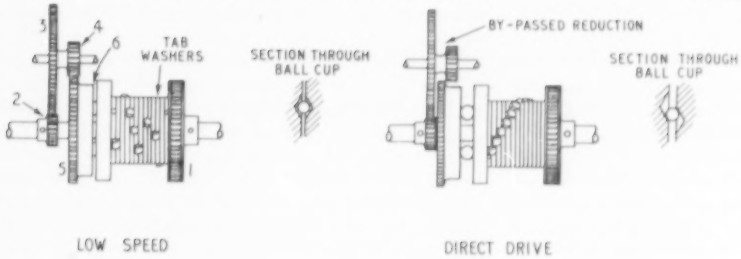
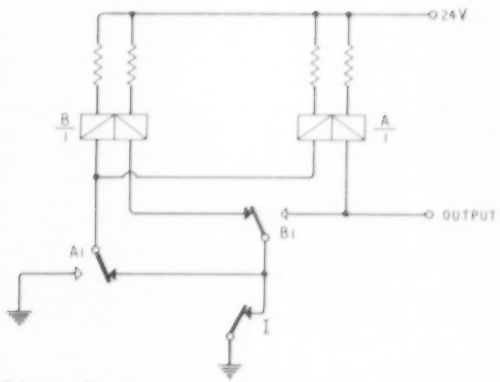


Fig. 27 (left). *By-pass mechanism.*

Fig. 28 (above). *Flash-back mechanism.*

Fig. 29 (below). *Basic scaler circuit.*



Binary Scaler

Because of the policy of employing unstabilized power supplies in the system, it was felt that a scaling circuit using valves might be unreliable. Moreover, the high counting speed afforded by a valve is quite unnecessary here, and so a simple relay scaler was considered to be well worth the introduction of a 24-V supply.

This circuit employs miniature high-speed relays with double-wound armatures and single change-over contacts. The system is basically that of Fig. 29 where I is an independent input contact. The phasing of the coils is such that if both windings are energized, the relay is not operated. The input relay (in this case the lock-on relay) and the two scaling relays together have eight possible states, and in the following table these states are listed in order of transition. The prime denotes the prefix 'not'.

Table of Relay States

	Contact	Armature	
1	I′ A′ B′	A′ B′	
2	I A′ B′	A B′	Inconsistent
3	I A B′	A B′	
4	I′ A B′	A B	Inconsistent
5	I′ A B	A B	
6	I A B	A′ B	Inconsistent
7	I A′ B	A′ B	
8	I′ A′ B	A′ B′	Inconsistent

In the first column, the condition of the contacts is postulated and, in the second, the condition of the armatures is determined by inspection of the circuit. It is seen that half of these are not self-consistent, but the removal of the inconsistency in the contacts leads naturally to the next state. Evidently the mechanism of a scaler is such that the input contact introduces inconsistent or unstable states into the circuit, and the wiring is such that this leads to a new stable state. Of the eight states only two can be stored without the help of the input relay or, what is more to the point here, there are two alternative states when the lock-on relay has operated, namely 3 and 7. In the main circuit diagram, Fig. 13, the lock-on relay is D corresponding to I in the table.

When the equipment is first switched on none of the relays is operated and this is state 1. The first operation of the lock-on relay shifts the scaler to state 3 and, if the loop phasing is suitable it stays there but, if not, the relay D opens and closes again, thus shifting the scaler to state 7. The phase-reversing relay C is so wired to the scaler that the relay is operated in state 7 but not operated in 3, so one or other of these conditions must be suitable for the control system.

Conclusion

The concept of the control plane, introduced for the purpose of describing the properties of a klystron and the behaviour of klystron control systems, has culminated in the construction of a map-painting device, with the aid of which a comprehensive a.f.c. system has been developed. Retention of the 'two-dimensional' outlook in discussing the performance enables one to escape such paradoxes as arise from attempting to associate a particular control with a particular task. It is sufficient to assert that the controls move jointly to annul both types of error. Yet, at the same time, the lack of inertia in the reflector control enables stability problems in two dimensions to be avoided. The velocity lag of the reflector due to the Miller integrator is not sufficient to cause appreciable deviation from the search track or frequency contour on account of the motor velocity, particularly when the extra gear reduction has been brought in. So that, provided the 'target' point is a fixed one, the joint control setting is confined to one or other of these loci.

The salient features of the a.f.c. system which has been described can be summarized as follows:

The fact that a definite relay operation marks the transition from search to follow, lends itself to the 'counting down' method of correcting control loop phase by trial and error. This method is foolproof in the sense that no circumstances can conspire to defeat its successful operation; only definite circuit faults can interfere with it.

The large amplitude sinusoid transient on the klystron reflector ensures easy mode finding, a good centring criterion, and leaves the controls unmodulated at the time when the receiver is being used.

Stabilized supplies are not necessary, because the significant properties of the klystron are under direct control. The supplies should, however, be adequately smoothed.

Finally, the by-pass gearbox concisely solves the difficulties of mechanical tuning by allowing an elementary constant speed on–off servo motor to be used without sacrificing setting accuracy. This mechanism effectively modifies the properties of the motor but does not demand any additional information from the control system.

The system described here is not restricted to a particular class of radar but the components of the discriminator circuits have to be chosen for the particular pulse length and repetition rate in mind.

Related Systems

In the light of this development work, it is possible to envisage more specialized methods of controlling klystrons for a.f.c. which may be attractive in particular circumstances. For example, in the system described above, the sinusoid amplitude is recommended to be about two mode widths peak-to-peak. If it is reduced to something less than a quarter of this magnitude, the system would degenerate to a mode-peak-finding servo with a restricted range of pull-in.

However, once pull-in had occurred it would be true to say that the klystron is frequency modulated with a deviation-time characteristic much the same in shape as the sinusoid itself.

Then, if the sinusoid could be accommodated within the time interval of the transmitted pulse, a similar f.m. transient would be available at the beat frequency (i.f.). This, together with a coherent time discriminator, could operate on a straightforward i.f. amplifier to produce a frequency-discriminator characteristic, thus dispensing with a conventional frequency discriminator altogether. Such an arrangement would have the drawback mentioned above but exhibit the following desirable properties:

The sinusoid would not occupy any of the usable part of the time scale. It could be triggered by the transmitter pulse.

The 'frequency discriminator' would be easier to design and produce than the Weiss or other types.

The discriminator would be coherent with the beat note sign and the binary scaler would be unnecessary.

Of course the use of such a system hinges on the practicability of accommodating the sinusoid 'inside' the transmitter pulse. With a typical i.f. of 30 Mc/s the pulse would need to be at least 2 μsec long to produce a carrier train long enough to frequency modulate.

Another variant of this method places no restriction on the transmitted pulse length but on the other hand severely restricts the rate of tuning-error correction. In this arrangement, the sinusoid is dispensed with altogether and the klystron reflector is modulated instead by a derivative of the transmitter pulse which has been amplitude modulated by a low-frequency sine wave (see Fig. 30). Using this waveform, the neighbourhood of the reflector potential is explored in discrete samples and error information can be extracted from the mode and frequency discriminators after one complete cycle of the sine wave. The time taken to accumulate a sufficient number of sample deviations is evidently proportional to the repetition period and so the system is best suited to radar equipments with high repetition rates. As such radars are essentially short range and therefore usually have short pulses, this a.f.c. system may be considered to be complementary to the previous one described.

Both of the above systems are being investigated. The simpler frequency discriminator and absence of the binary scaler promised an economy in components but, on the other hand, the reduced sinusoid amplitude requires a more sensitive mode-centre discriminator because the crystal-current variations at mode centre are comparatively small. Furthermore, it is desirable to use a separate crystal for mode-centring information when measurements are being performed during the transmitter pulse because the transmitter makes an unwanted contribution to the mixer crystal current at that time.

It may be necessary to augment this discriminator with some form of reflector scanning to restore the mode pull-in range.

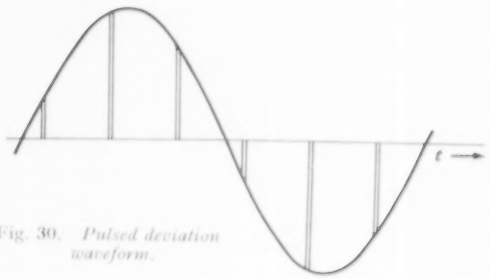


Fig. 30. *Pulsed deviation waveform.*

Acknowledgments

When the development of this system was commenced, the reports of S. Ratcliffe (now published[3]) were already to hand and did much to clarify the problem and unify the diverse opinions at that time. His recommendations constitute a guide to good practice in radar a.f.c. and no contemporary design can be said to be uninfluenced by them.

The method of interconnecting the error circuits and the klystron controls was proposed by V. J. Cox, who also introduced the innovation of the wide amplitude sinusoid for mode centring. These features are the basis of the system.

The circuit of the binary scaler is that described by B. E. Swire[4] and its use here to combat instability constitutes an elementary example of a type of circuit discussed at length by Dr. Ross-Ashby and classified as ultra-stable[5].

The step cam simulator in the gearbox is an invention of S. Zillwood Milledge[6].

Much of the initial development and experimental work was carried out by R. Beaven.

To these, and to the Directors of E. K. Cole Ltd., who granted permission to publish the paper, the writer is duly grateful.

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BACK-SCATTER IONOSPHERIC SOUNDER

Single-Station Equipment for Oblique Incidence Propagation Studies

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(Official communication from D.S.I.R. Radio Research Station, Slough)

SUMMARY.—*High-frequency ionospheric propagation can be studied by observation of scattered echoes received by way of the ionosphere from irregularities on the earth's surface. The paper describes an equipment developed for propagation investigations by this method. The design of a suitable 150-kW pulse transmitter which can be simply tuned to any frequency in the band 10–27 Mc/s is discussed. To confine the observations to certain chosen directions, three-wire rhombic aerials are employed, the same aerials being used for both transmission and reception by employing a tunable transmit–receive switch. A receiver, which is of a type developed for ionospheric-absorption measurements, provides an output suitable for the presentation of the received echoes on a normal time-base display. To provide permanent records, photographs are taken of this display and, in addition, continuous range-time (p't) recordings are made which show the variation with time of the echo pattern. Echoes are received from ranges up to about 3,000 km by single-hop F2-layer propagation and up to 12,000 km by multi-hop propagation.*

The same transmitter and receiver are also used in conjunction with a continuously-rotating Yagi aerial and a plan-position indicator. By the use of a speeded-up ciné film technique, the changes occurring in this display during a complete day may be seen in a few minutes, giving the observer a demonstration of the changing pattern of ionization in the ionospheric layers as the sun's illumination moves over the area surveyed.

1. Introduction

IN the planning and day-to-day operation of long-distance h.f. communication, a means of measuring directly the maximum-usable frequency for propagation between transmitting and receiving stations is very desirable. A new method, by which such measurements may be made at a single station without the co-operation of distant observers, has recently come into prominence. In this technique, pulses of radio waves are radiated from a directional aerial along the path to be studied. As in normal h.f. communication, energy which is emitted below a certain critical angle of elevation is reflected by the ionosphere and returned to the earth's surface at distances from the transmitter greater than the skip distance. A small part of this incident energy is returned to the vicinity of the transmitter by scattering from irregularities on the ground (Fig. 1). This scattered energy can be detected at the transmitter in the form of weak echoes of the transmitted pulse, the time delay of the earliest echoes to arrive being a measure of the oblique range to the edge of the skip zone. If the height of the reflecting layer is known, it is possible to use the oblique range to calculate the skip distance. The maximum-usable frequency for the chosen path is then the frequency for which the measured skip distance is the same as the distance to the receiving station.

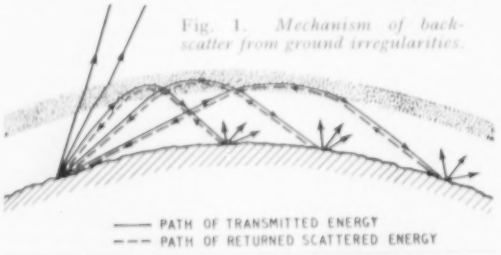


Fig. 1. *Mechanism of back-scatter from ground irregularities.*

MS accepted by the Editor, September 1955

The present paper discusses the requirements for a pulse sounder for making such measurements and describes the design of a particular equipment developed at the Radio Research Station. The theory of propagation of scattered echoes and the methods of analysing results have been studied in two earlier papers [1,2] and will not therefore be discussed in detail here.

2. General Considerations

The essential components of a pulse sounder for back-scatter observations are a pulse transmitter, directional transmitting and receiving aerials, and a pulse receiver with a suitable time-base display. A simplification of the aerial system is made possible by the use of a single aerial arranged for common transmitter–receiver operation.

The most flexible system would have a transmitter and receiver tunable to any desired frequency in the h.f. communication band and coupled to an aperiodic rotatable aerial. With such an arrangement it would be possible to measure the m.u.f. for communication to any point within the range of the equipment, or alternatively to study the large-scale movements and changes taking place in the ionospheric layers over the area. However, the design of a suitable aperiodic rotating aerial presents considerable engineering

problems and in the present system some compromise has had to be made between the requirements for variation of direction and variation of frequency.

The aerial system, which is arranged for common transmitter–receiver operation, comprises four aperiodic aerials of fixed azimuths and a rotating aerial with a fixed operating frequency. The aperiodic aerials are used over the whole of the frequency band 10–27 Mc/s covered by the associated pulse transmitter and receiver. Permanent records of the echoes obtained are made by taking photographs of the normal time-base display while, to study the variation of echo range with time, continuous recording on moving film is adopted.

For displaying the echoes obtained with the rotating aerial, use is made of the plan-position indicator, first suggested for application to back-scatter research by Newbern Smith[3] and later developed by Villard and Peterson[4] and de Bettencourt[5]. This is a very convenient method of presentation, especially if an outline map is superimposed on the face of the c.r. tube, since the zones of the earth from which echoes are being returned, and to which radio communication is therefore possible, are at once obvious.

To assist in the study of the diurnal changes in size and position of these 'illuminated' zones, photographs are taken on ciné film at intervals of about a minute and after processing are projected at 16 frames a second. The showing of such a film gives in a few minutes a very clear demonstration of the changing pattern of ionization as the sun's illumination moves across the area surveyed.

In the following account of the development of the back-scatter sounder, the transmitter, receiver and fixed aerials are considered first, since together they comprise a complete equipment for the study of propagation in certain chosen directions. The rotating aerial, plan-position indicator and associated ciné camera are then discussed together since their design is interrelated.

3. Transmitter. Factors affecting Design

The factors governing the design of a transmitter for back-scatter work are the peak power in the pulse, the pulse width and the pulse repetition frequency. There is, in addition, the need for a simple tuning arrangement so that the frequency can easily be changed.

The choice of pulse repetition frequency (p.r.f.) is governed by the range from which echoes are received which, for high-power transmission, is frequently 5,000 km and under favourable conditions can extend up to 12,000 km corresponding to a total echo time delay of 80 ms. In this equipment the normal p.r.f. is 25 c/s (40 ms period) with an alternative of $12\frac{1}{2}$ c/s (80 ms period) available when required.

The duration of the radiated pulse need not be less than 100 μs for back-scatter measurements since accurate range discrimination is not needed and, in addition, pulses shorter than this would generate too much interference to stations working on adjacent frequencies. A receiver bandwidth sufficiently wide for reception of short pulses would also increase interference from adjacent stations and make detection of echoes difficult.

In view of the poor signal-to-noise ratio and limited range of the echoes obtained in preliminary experiments, in which a 10-kW transmitter radiating 140-μs pulses was used, it was decided to design for a power exceeding 100 kW. An alternative method of improving the signal-to-noise ratio of back-scatter echoes adopted by Villard and Peterson[4] is to use a low-power transmitter (1–10 kW) radiating long pulses (1–2·5 ms). This system leads to a simpler equipment which is probably sufficiently accurate for operational use in the measurement of skip distance. It has, however, been found that the higher range discrimination obtained with the short-pulse equipment is preferable for research work and has led to some interesting results concerning the time variation of the scatter pattern (see Section 8).

3.1. *Transmitter. Main framework*

The transmitter was designed round the power supply and valves of a radar transmitter, R.A.F. type MB2[6], which originally generated 10-μs pulses of 300-kW peak power at a p.r.f. of 25 c/s on either of two predetermined frequencies. The valves in this transmitter had been developed from c.w. types in the early days of radar and consequently had a higher anode dissipation than was necessary for 10-μs pulses. This fortunate circumstance made it possible to obtain a power output of 150 kW for 100-μs pulses without exceeding the permissible dissipation. With later radar valves, the small allowable dissipation would have limited the power output while, with c.w. valves, the high emission necessary (70 A for the output valves) could only have been obtained with very large water-cooled types having unnecessarily high maximum dissipation.

Fig. 2 shows the r.f. compartments of the modified transmitter. Only minor changes were necessary to the power supply compartments in the lower part, but the pulse generator and radio-frequency compartments were stripped completely, apart from the valves and valve mountings.

The method of modulation, by interruption of a bias current through the oscillator grid leak, was left unchanged, but new pulse generating and

monitoring units were built and mounted on a rack beside the transmitter. These circuits follow conventional modern practice and include provision for measurement of the duration of the radiated r.f. pulse.

3.2. *Transmitter. R.f. circuits*

To satisfy the requirement for a simple tuning arrangement to cover the frequency range 10–25 Mc/s, new r.f. circuits were designed to replace the existing tuned-anode–tuned-grid arrangement with its multiplicity of controls and restricted coverage.

Variable inductors were adopted for tuning in preference to variable capacitors because of the larger frequency range ($2\frac{1}{2}$:1) achievable with the former, and the smaller volume occupied by the complete tuned circuit assembly. The complete r.f. circuit of the transmitter is shown in Fig. 3, and is seen to consist of a Colpitts oscillator coupled to a tuned amplifier.

In the oscillator tuned circuit, a variable inductor is employed having a single rotating wiper which is electrically connected to one end of the coil, and which can be wound inwards from that end of the coil. Although the tuned circuit is balanced with respect to earth, no undesired unbalance is introduced by this unsymmetrical wiper arrangement since any unbalanced stray capacitance introduced is small compared with the balanced capacitance to earth provided by the potentiometer chain.

The power amplifier, however, is tuned by a variable inductor having two wiper arms which are wound in with opposite directions of rotation from the two ends of the coil. This is necessary to keep the coil electrically symmetrical, since the aerial-coupling coil is moved in and out of a space between the centre turns of the tuning inductor. To provide the required insulation, the coupling coil is moulded in synthetic resin, the complete moulding being free to slide between vertical guides. The vertical position of this coil, and therefore the degree of coupling, can be controlled by a knob on the front panel of the transmitter operating through a cam mechanism.

In order to achieve the close coupling required between the tuned circuit and the aerial coil, this coil is tuned by a shunt variable capacitor. The Q factor of this circuit is very low, varying between 1 and 2 over the band, so that the tuning is not critical. In practice three settings of this control are used for the frequency ranges 10–12, 12–16 and 16–27 Mc/s. When the transmitter is operating into a feeder impedance of 600 ohms, no adjustment of the coupling coil is required, the

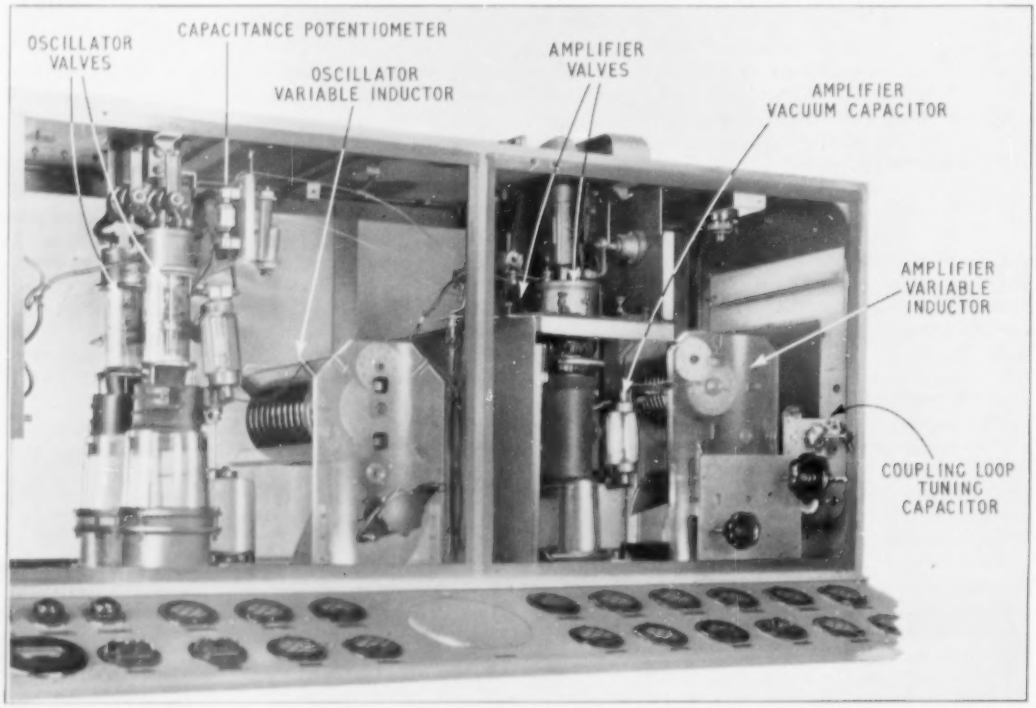
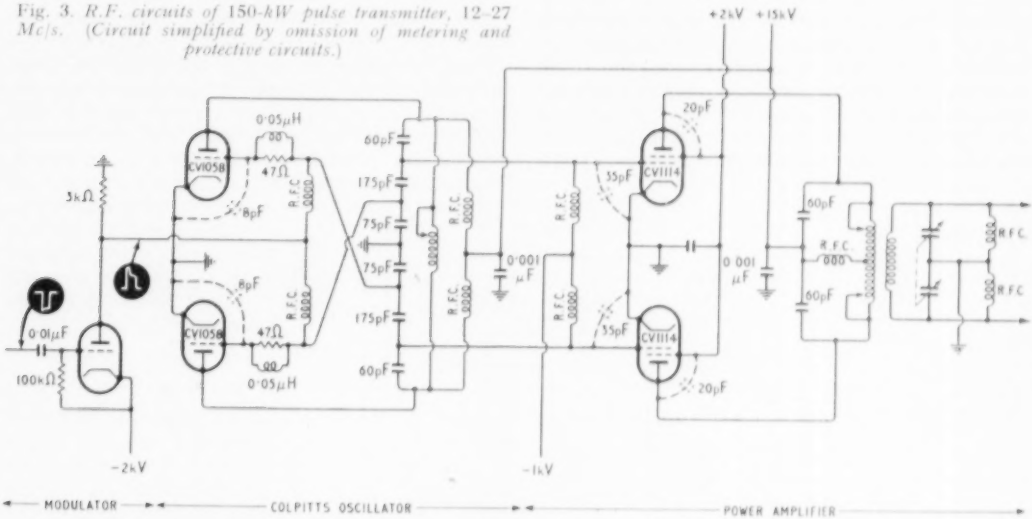


Fig. 2. *Transmitter–R.F. compartments.*

coil being left at the position where it is coaxial with the tuning coil.

The transmitter is normally operated with a d.c. pulse power input to the oscillator of 20 kW, and to the amplifier of 240 kW, the h.t. to both stages being 10 kV. Under these conditions a peak r.f. output of 110 kW is obtained.

Fig. 3. *R.F. circuits of 150-kW pulse transmitter, 12–27 Mc/s. (Circuit simplified by omission of metering and protective circuits.)*



When the frequency is varied, the power remains approximately constant at 110 kW from 10 to 23 Mc/s; above this range it drops rapidly, falling to 70 kW at 25 Mc/s. It is found that the frequency stability of the transmitter is high. This is attributed to the high Q factor of the oscillator-tuned circuit (50–200), the avoidance of phase shift in the feedback network and the stability of the inductors and vacuum capacitors.

4. Rhombic Aerials

For the study of propagation in directions from Slough which are of particular interest, two horizontal rhombic aerials have been erected. Open-wire feeders are provided from both ends of each aerial to the building in which the transmitter is situated; with this arrangement the direction of shoot can be reversed by interchanging the connections of the transmitter and terminating resistance. The aerials are identical in size and shape, each having a semi-side angle of 70°, a length of side of 64·6 metres and a height above ground of 13·7 metres.

A peak r.f. voltage of 13·5 kV is developed across the 600-ohm feeders at 150 kW transmitted power, and adequate spacing and insulation to withstand this voltage has therefore been necessary at all points on the system. By the adoption of three-wire construction for the rhombic conductors, a constant impedance–frequency characteristic has been achieved and standing waves minimized.

The calculated azimuthal beamwidth of the aerials for common TR operation at 17 Mc/s is $\pm 5°$ to half-power points. Maximum radiation in the vertical plane takes place at an angle of elevation of 17°, with half-power response at 12° and 22°.

No experimental measurements of the directivity of the aerials have been made with the exception of a brief investigation of the relative response in forward and backward directions. The measured value of this back-to-front ratio was 15–20 dB for one-way propagation and 30–40 dB for echo reception. Comparisons of the echoes received with the rhombic aerials and those observed with the rotatable aerial show satisfactory rejection by the rhombics of echoes from directions different from that of the main lobe, but investigation of this is still in progress.

5. Transmit–Receive Switch

To make possible the use of a common aerial for transmission and reception a transmit–receive (TR) switch capable of protecting the receiver during the transmitted pulse has been developed. Use could not be made in this application of the conventional radar type of TR switch consisting of short-circuited $\lambda/4$ transmission lines, since operation at a number of frequencies was desirable. To overcome this difficulty, a balanced artificial line is employed, consisting of one prototype low-pass filter section with fixed inductors as the series arms and a variable capacitor as the centre shunt arm. The complete unit is

connected between the transmitter terminals and the receiver, the aerials being connected directly to the transmitter (see Fig. 4).

An argon-filled spark gap (CV 1507) of the type described by Banwell[7], is used as the switching element and is shunted across the variable capacitor. When the transmitter is radiating, the spark gap conducts, the receiver is short-circuited and the impedance of two right-hand 17-μH coils (Fig. 4) is shunted across the transmitter terminals. Since this impedance is 2,500 ohms inductive at 10 Mc/s, and still greater at higher frequencies, the only effect of the TR switch on the transmitter is to alter the tuning of the coupling coil very slightly. As this is in a low-Q circuit, the effect is negligible.

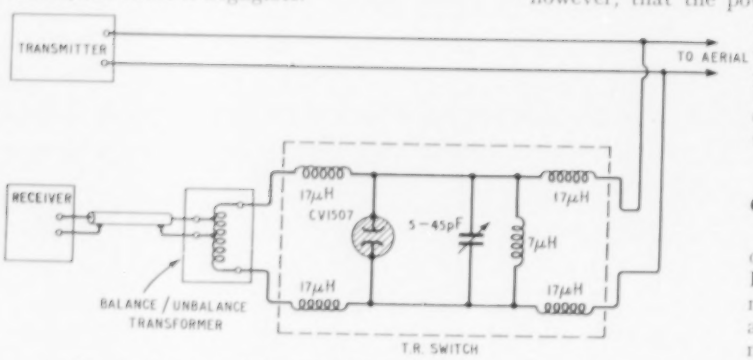


Fig. 4. *Transmit–receive switch circuit.*

In the interval between transmitted pulses, the spark gap is non-conducting and, by adjustment of the tuning capacitor, the network may be made electrically $\lambda/2$ long. This is achieved in practice by tuning the capacitor until the echoes observed with the receiver are of maximum strength. Since the network is $\lambda/2$ long and symmetrical, the impedance seen by the aerial is equal to the impedance of the receiver (neglecting losses), and the matching of the receiver to the aerial is unaffected. A wideband transformer matches the 100-ohm unbalanced receiver impedance to the 600-ohm balanced feeder.

The frequency coverage of the switch is governed by the ratio of maximum to minimum capacitance of C. This ratio has been made high by employing a capacitor with two stators and a butterfly-section rotor; it is used in a balanced condition with the rotor at earth potential although not physically connected to earth. An increase of the effective capacitance ratio is achieved by the inclusion of the 7-μH inductance which reduces the minimum capacitance from 5 pF to zero at 25 Mc/s, at the same time diminishing the maximum at 10 Mc/s from 45 pF to 11 pF. With these values the switch can be tuned satisfactorily over the range 10–25 Mc/s, the insertion loss in the receiving condition being 1 dB at 10 Mc/s, 3 dB at 20 Mc/s and 5 dB at 25 Mc/s.

It will be noticed that, in the receiving condition, the coupling coil of the transmitter is still connected across the aerial feeders. When the transmitter is quiescent, the power-amplifier tuned circuit has a high Q and a tuning point can be found, close to the resonant frequency of the circuit, for which the coupled reactance appearing in series with the output circuit cancels out the coupling-coil reactance. A low resistance is thus thrown across the aerial feeder and the received echoes are severely attenuated. It is found, however, that the power-amplifier tuned circuit can be detuned slightly to avoid this condition without affecting the power radiated by the transmitter, since the Q of the tuned circuit when transmitting is low.

6. Receiver

The receiver is of a type developed at the Radio Research Station for the measurement of ionospheric absorption by a pulse technique. A detailed account of the design has appeared elsewhere[8], and accordingly only a few features of interest will be discussed here. The receiver is arranged to deliver push-pull deflection voltages to the Y-plates of a cathode-ray tube, and to have a high degree of linearity between input and output voltages for the range of amplitudes which can be conveniently observed on the screen of the tube. In addition, the sensitivity of the receiver is unaffected by the presence of pulse or keyed c.w. transmissions strong enough to cause overloading. As a result, when the receiver is used with the transmitter and TR switch already described, no decrease of sensitivity after the pulse can be detected.

The Y-plates of the c.r. tube are connected direct to the load resistance of two diode detectors giving positive and negative d.c. outputs respectively of up to 300 volts. This d.c. coupling method eliminates any vertical drift of the time-base when keyed c.w. interference is present and results in a very satisfactory display for visual observation.

An overall gain of 170 dB is achieved, which is sufficient to give full deflection on the c.r. tube screen for thermal noise in the first tuned circuit. The bandwidth is 12 kc/s between half-power points.

7. Time-Base Generator and Display Circuits

The arrangements for synchronizing the transmitter pulse, time-base and display range-markers are shown in the schematic diagram of Fig. 5. 25-c/s or 12½-c/s square waves derived from a stable oscillator by frequency division are used to synchronize the circuits.

Two cathode-ray tubes are available for display, one with a green screen for visual observation and the other with a blue screen for photography. The waveforms applied to the two tubes are identical and are indicated in Fig. 5. A 25-c/s time-base is normally used for visual observation since the flicker of a 12½-c/s time-base is found to be fatiguing to the eyes. It is, however, essential to use 12½-c/s if any echoes of greater time delay than 40 ms are present.

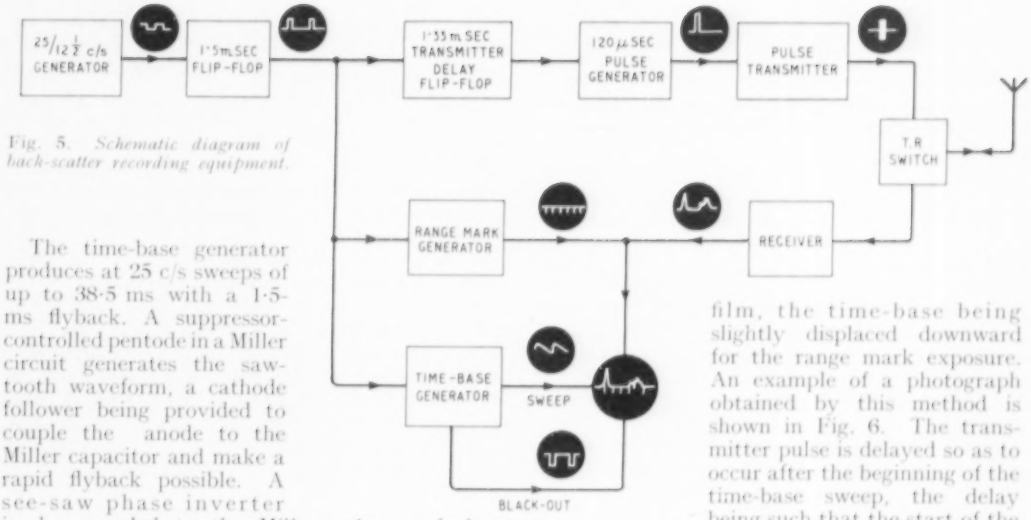


Fig. 5. *Schematic diagram of back-scatter recording equipment.*

The time-base generator produces at 25 c/s sweeps of up to 38·5 ms with a 1·5-ms flyback. A suppressor-controlled pentode in a Miller circuit generates the saw-tooth waveform, a cathode follower being provided to couple the anode to the Miller capacitor and make a rapid flyback possible. A see-saw phase inverter is d.c. coupled to the Miller valve, and d.c. coupling to the X plates of the c.r. tubes is adopted. By this means, any drifting of the position of the time-base when the sweep speed is changed is eliminated. A negative-going black-out pulse is taken from the screen of the Miller valve and applied to the grids of the c.r. tubes. Several alternative sweep durations are available (see Table 1) so that back-scatter echoes of various ranges, meteor echoes or pulse transmissions from distant stations may each be observed on a time-base of convenient length. Ranges given in the table are strictly 'equivalent free-space ranges' and assume uniform free-space velocity of propagation. The corrections necessary for accurate work are discussed in Section 8.

Range markers are provided either at 50-km intervals with every fourth mark enlarged, or at 200-km intervals with every fifth mark enlarged. A 3-kc/s Hartley oscillator provides the timing source, the oscillation being interrupted during the flyback of the time-base and started again at the beginning of the sweep so as to provide a stationary pattern of markers. The 50-km (330-μs) markers are obtained from the 3-kc/s sine wave after suitable shaping, while the 200-km (1·33-ms) and 1,000-km (6·67-ms) markers are obtained from the same waveform by frequency division in two phantastrons. After suitable mixing and amplification, the markers are combined with the signal from the receiver so that the range marks appear superimposed on the echo pattern as downward-going pulses. For photographic recording it is found that analysis of the records is greatly eased if separate exposures of the range marks and the echo pattern are made on the same film, the time-base being slightly displaced downward for the range mark exposure. An example of a photograph obtained by this method is shown in Fig. 6. The transmitter pulse is delayed so as to occur after the beginning of the time-base sweep, the delay being such that the start of the pulse is coincident with the first 200-km marker.

A further type of photographic record which has yielded very useful information is the continuous

TABLE 1

Sweep Duration (ms)	Range (km)	Type of Echo Observed
2·3	350	Meteors. Pulse transmissions from distant stations.
6	900	Meteors. Back-scatter reflected by sporadic-E ionization.
17	2,600	Single-hop back-scatter.
36	5,400	Single- and double-hop back-scatter.
55	8,200	Multi-hop back-scatter.
75	11,200	Multi-hop back-scatter.

range-time recording ($p't$) of which an example is shown in Fig. 7. To obtain such records a linear time-base, brilliance-modulated by the signal, is photographed, the sense of the modulation being such that an echo blacks out the spot. Film is drawn through the camera in a direction at right angles to the image of the time-base line. To provide range and time calibrations, a clock is arranged to change the modulation from signal to range markers for a few seconds every 10 minutes. A further time identification is given by omitting this calibration at the hour.

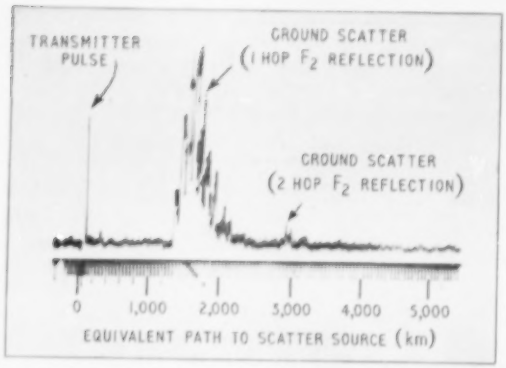


Fig. 6. *Typical scatter pattern for winter noon.* 1110 *GMT,* 13.11.1951 *aerial, rhombic* 80° *E of N; frequency* 18·5 *Mc/s.*

8. Discussion of Results obtained with fixed Directional Aerials

Although it is not proposed here to analyse in detail the results obtained with the present equipment, a brief discussion of the photographs given in Figs. 6 and 7 may be of interest. Fig. 6 is a typical example of the echo pattern seen on a linear time-base display when propagation is taking place by way of the F layer alone. The lack of echoes at a shorter range than 1,350 km is explained by the skip phenomenon; since no energy was reflected back to the earth's surface in the skip zone, no energy could be returned to the transmitter by scattering. Strong echoes were, however, received from just beyond this skip zone, since the energy reaching this region was enhanced by skip focusing.

The enhanced intensity of the echoes from the neighbourhood of the skip distance is a fortunate circumstance for skip-distance measurement, since echoes from this area can be distinguished above background noise at times when echoes from greater distances are too weak to detect. A measurement of the skip distance can thus be made whenever back-scatter echoes can be seen. To obtain the skip distance from the equivalent free-space range of the leading edge of the scatter pattern (1,350 km in Fig. 6) a correction must be applied to allow for the path obliquity and the low velocity of propagation in the layer. This correction, which must be subtracted from the free-space range, is of the order of 150 km for distances greater than 1,000 km; values for various layer heights have been calculated and given in graphical form elsewhere[1].

A feature of the echo pattern of Fig. 6 which should be mentioned is the group of 'second-hop' echoes at 2,800 to 3,200 km range. This is due to energy which on the outward journey suffered two reflections from the ionosphere with an intermediate reflection at the ground. These echoes were also enhanced in intensity by the skip focusing phenomenon referred to previously.

An example of a continuous range-time ($p't$) record of back-scatter is given in Fig. 7. The observation was made at Slough in February 1955 at a frequency of 19 Mc/s using a rhombic aerial directed towards the south. The scatter is indicated by the black areas, the echoes from the first skip focusing region being recorded at ranges of 2,000 to 3,000 km before finally fading out at 1830 GMT. Other multiply-reflected groups of scatter can be seen at greater ranges, but it will be noticed that the ranges are not necessarily multiples of the range of the first group, and that one

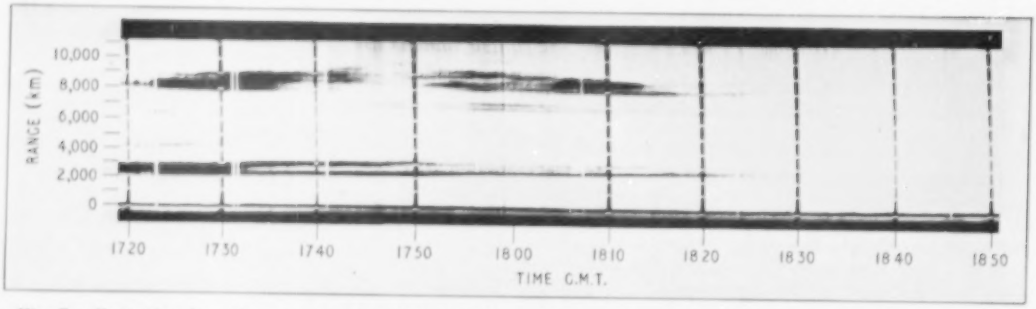


Fig. 7. *Example of continuous range-time* ($p't$) *recording of back-scatter. (Single-hop and multi-hop echoes can be distinguished.) Rhombic directed* 139·5° *E,* 21*st February* 1955; *frequency* 19 *Mc/s.*

'multiple-hop echo' persists after the first-hop echo has faded out. From experimental information such as this, it appears that the mode of propagation of energy over long distances can be very complex. An investigation of these phenomena is in progress.

9. Back-Scatter Plan-Position Indicator. General Considerations

In principle, the back-scatter plan-position indicator (p.p.i.) is identical with that used in radar, the differences being due to the very much greater range of the echoes and the very much longer wavelength. As a consequence of the long wavelength, the rotating aerial system needed to get the angular resolution possible with a radar aerial would be prohibitively large, and an inferior performance has to be tolerated. The present Yagi aerial has a beamwidth for common TR working of 40° between 3-dB points, which does not give sufficient resolution to register correctly ionospheric changes taking place over a small range of azimuth. In spite of this limitation, however, it has provided much useful information about geographical and temporal changes in the ionosphere.

The long range from which the echoes are received imposes an upper limit on the speed of rotation of the aerial and time-base because of the low p.r.f. necessary (25 c/s for a range of 6,000 km from Section 3). For a satisfactory display, it has been found that at least one radial line per degree of rotation is needed, so that at 25 c/s the time of rotation must not be less than 40 ms $\times$ 360 = 15 secs. The maximum speed of rotation is thus 4 r.p.m. at 25 c/s, or 2 r.p.m. at 12½ c/s.

9.1. *Rotating Aerial*

The type of directional aerial which was adopted for the p.p.i. equipment was a four-element horizontal Yagi. For its small size, about $\lambda/2$ square, and simple construction, such an array gives a directivity that could only be appreciably improved upon by a considerably larger and more complex aerial. A disadvantage is that its use is limited to within about $\pm 2\%$ of the designed frequency but, for the present purpose, it was decided to accept this limitation, and an operating frequency of 17 Mc/s was chosen. Scatter observations at this frequency give information about daytime propagation for lengths of hop of the order encountered in long-distance communication (1,500–3,000 km).

The aerial, illustrated in Fig. 8, occupies an area in the horizontal plane of 30 ft by 23 ft and is mounted on a tower 30 ft high. The driven element lies across the centre of a rotating wooden framework and consists of two telescopic rods which can be adjusted in length to give a resistive input impedance at the centre of the element. A reflector is spaced $0{\cdot}2\lambda$ behind and two directors $0{\cdot}1\lambda$ and $0{\cdot}2\lambda$ in front of the driven element. The two directors and the reflector each have at their centres a tuning stub with a short-circuiting bar, the position of which can be adjusted manually from the centre of the framework by means of an insulated rod. With this arrangement, it is possible to adjust the electrical length of the elements to give the optimum gain consistent with a good back-to-front ratio.

The wooden framework supporting the elements rotates about a central spindle which is mounted in ball-races fixed to the tower. By means of a

Fig. 8. *Rotating Yagi array.*

sliding key the spindle is constrained to rotate with the framework, although the weight of the framework is not supported by the spindle. The same spindle passes downwards into a metal box where it drives a magslip transmitter through a 30:1 step-up gear ratio. This magslip is electrically connected to a magslip receiver which rotates the time-base coils of the c.r. tube indicator through a 30:1 step-down gear ratio.

The weight of the rotating wooden framework is carried by four wheels which run on a 6-ft diameter circular track of 2-in. angle iron, the track being divided into four insulated arcs to reduce any distortion of the aerial polar diagram. The framework is caused to rotate by a fifth wheel driven by a d.c. motor through a gearbox. The pressure of the wheel on the track is increased by a helical spring to provide the necessary amount of friction.

R.f. power transfer from the 600-ohm feeder to the rotating centre element of the aerial (impedance 25 ohms) is accomplished by inductive coupling between two coils mounted coaxially with the centre spindle. The inner, fixed, coil consists of four turns of 3/8-in. wide strip wound in a helix 5½ in. in diameter and is connected across the 600-ohm feeder with a 20 – 30 pF shunt tuning capacitor. The rotating coil is 7½ in. in diameter and consists of a single turn of 3-in. wide strip tuned by a 400-pF 'pot' type mica capacitor. An electrostatic screen, in the form of a cylindrical shell having a gap in its circumference, is incorporated between the primary and secondary coils to prevent capacitive coupling. Connection between the aerial and the single-turn coil is by a short length of balanced transmission line of 25 ohms characteristic impedance made of two strip conductors 1 in. wide insulated from each other by a 1/8-in. thick sheet of polystyrene.

The impedance of the driven element of the Yagi aerial can vary between 20 and 50 ohms when the lengths of the directors and reflector are changed[9] and this impedance is transformed to 600 ohms by the coupling coils. A variation of the transformation ratio over the required range is effected by moving the outer, single-turn coil axially, thus varying the inductive coupling between the coils.

To measure the performance of the aerial after erection, a field was set up by means of an elevated horizontal loop transmitter at 200 metres range, and the voltage appearing across the 600-ohm coil terminals was measured. Adjustments were then made to the reflector and director tuning stubs to obtain the best compromise between high gain in the forward direction and a good front/back ratio.

The measured azimuthal beamwidth for 3-dB drop in response, after squaring the polar diagram for common TR working, was 42°, and the back-to-front ratio 38 dB for the same conditions. No measurements have been made in the vertical plane, but the angle of elevation for maximum radiation obtained by calculation is 25°.

The measured gain of the array relative to a half-wave dipole was 6 dB, this measurement being obtained by comparing the Yagi pick-up with that of a reference dipole fixed temporarily to the tower before the erection of the Yagi.

9.2. *P.P.I. Display*

For visual observation of a p.p.i. display, it is necessary to use a c.r. tube with an afterglow of sufficient duration to make the whole display visible simultaneously. As discussed in Section 9, a time-base rotation speed of not greater than 2 r.p.m. is necessary with a back-scatter p.p.i. when operated at a p.r.f. of 12½ c/s, so that an afterglow of at least 30 secs is required. The c.r. tube used (Type CV464) has a 9-in. diameter screen with a fluoride coating giving a suitable long persistence orange afterglow with no initial flash. The tube has electrostatic focusing and requires magnetic deflection by external coils. Magnetic deflection is necessary since the high beam current needed for a satisfactory p.p.i. display precludes the use of electrostatic deflection plates in the c.r. tube.

A rotating time-base is produced by passing a sawtooth current waveform through a pair of deflecting coils mounted around the neck of the c.r. tube, the coils being rotated in synchronism with the aerial using the magslip synchronous link referred to in Section 9.1. The rotating-coil assembly forms part of an ex-Service p.p.i. console which also includes display centring facilities.

A time-base generator of the type described by Williams, Howell and Briggs[10] is used, suitable modifications being included to permit operation with a flyback time small compared with the sweep duration.

9.3. *Presentation of the Signal*

For back-scatter echoes it has been found preferable to show the echoes on the p.p.i. as dark patches on a light background; with the normal method of operation, it was found that the time-base was brightened by interference from other transmissions and the afterglow persisted for several revolutions of the display. With the system adopted this was overcome and, in addition, the sensitivity for weak echoes was improved.

Range markers are applied to the time-base as brightening pulses which produce circles at radial intervals of 200 km or 1,000 km when the p.p.i. is operating. For convenience, an outline map of the area covered is drawn on the face of the c.r. tube so that the location of the ground areas from which the scattered echoes are being returned can be seen.

The circuits which combine the signal, range

marker and flyback-suppression waveforms into a form suitable for application to the c.r. tube have been designed to retain the original d.c. level of the signal independent of the shape of the echo pattern or of the amount of interference present. By this means the brightness of the background remains constant in the spacing intervals of telegraph interference, the time-base being blacked out only during the instants when the interfering carrier is being received. Interference does not affect the brightness of the range circles.

If care is not taken to maintain the d.c. level of the signal, the background illumination in the spacing intervals of an interfering c.w. transmission is brightened or even defocused, causing mutilation of the display; this is unsatisfactory for visual observation and very serious for photographic recording.

Two features of the circuit, which are made necessary by the low repetition frequencies of 25 c/s and 12½ c/s required for scatter work, are the use of long time-constants in the coupling circuits to the c.r. tube, and the achievement of a high degree of smoothing in the negative e.h.t. supply. The latter is required because of the brilliance modulation produced by the ripple currents which flow through the necessarily large grid and cathode coupling capacitors. The brilliance modulation produced by any residual ripple in the supply is reduced by the use of identical coupling capacitors for the grid and the cathode. Elimination of brilliance modulation of the display by mains ripple is more important for photography than for visual observation because of the smaller range of luminosity which can be registered by photographic emulsions.

9.4. *Photography of the P.P.I. Display*

Permanent records of the p.p.i. display patterns are obtained by photography of the display, a ciné camera being employed to provide the 'speeded up' ciné films referred to in Section 2.

For this purpose, the aerial and time-base are rotated at 1 r.p.m., and a single frame of 16-mm ciné film is exposed for each rotation. The shutter of the camera is closed only during the instant at which the film is wound on, so that the complete 'paint' produced by one rotation of the time-base is photographed. To provide a record of the time, a watch is illuminated for one second during each exposure. After processing, the film is projected at the normal speed of 16 frames a second, so that the changes occurring during 24 hours may be seen in 1½ minutes.

Another type of photographic record which has proved useful for some purposes is an azimuth-range plot in Cartesian co-ordinates. This is produced by a similar technique to the range-time recordings referred to in Section 7. The difference here, however, is that the film is kept stationary while a horizontal brilliance-modulated time-base is moved slowly in a vertical direction, the vertical displacement at any time being proportional to the angle through which the aerial has turned. This type of record is made with the same c.r. tube and camera as is used for normal time-base display photography.

9.5. *Discussion of Results obtained with the P.P.I.*

As an example of the results obtained with the back-scatter p.p.i., some still photographs taken from a ciné film of the kind described in Section 9.4 are shown in Fig. 9. The frequency of transmission was 17 Mc/s, which was above the maximum frequency for which oblique rays could be reflected by the night-time ionosphere, but was low enough for reflection to take place from those areas of the ionosphere illuminated by the sun.

In the morning, back-scatter was first received from ground areas to the south-east after sunrise in the F layer of the ionosphere. As the day progressed, scatter was also received from the south, west and north. In the photograph for 1000 UT, a complete ring of scatter can be seen with a minimum range (approximately equal to the skip distance) of 1,500 – 2,000 km. Propagation on 17 Mc/s by way of the F layer was thus possible in all directions at this time. A second arc can also be seen to the south outside the main ring of scatter. This indicates that energy was being returned from ground areas illuminated by two-hop rays; that is, rays which on their outward and inward journeys were reflected twice from the layer with an intermediate reflection at the ground.

In the photograph for 2100 UT, scatter is being received by way of the F layer only from the west, the rest of the area being in darkness. The short-range echoes recorded from 1240 UT to 2100 UT represent ground scatter reflected by localized clouds of 'sporadic-E' ionization.

Within the first ring of scatter is the skip zone to which no communication is possible by normal ionospheric reflection. Between the first and subsequent rings are zones where the field strength is low, but not necessarily so low that communication is impossible; to receive scatter from such ranges, higher power, longer pulses, or improved aerials and receiving systems would be required.

10. Conclusions

The equipment described in this paper was developed for an experimental investigation of h.f. propagation by the back-scatter technique. The 150-kW pulse transmitter has been used in conjunction with the rhombic aerials to study propagation over the band 10–27 Mc/s in certain fixed directions. Ground scatter propagated by

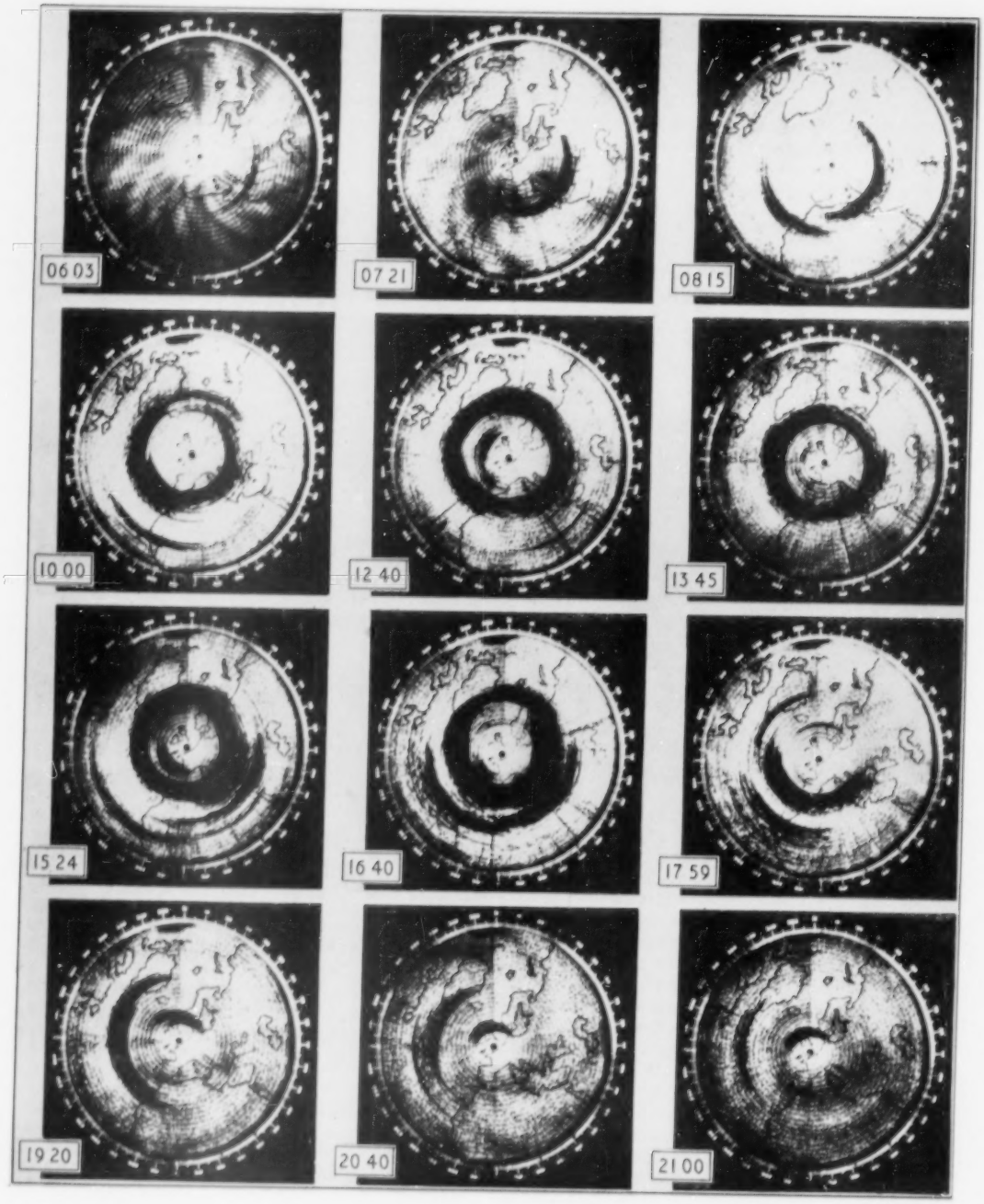


Fig. 9. *Variation of back-scattered echoes over the period* 0600–2100 *UT, 7th October* 1954.

way of the F, E and Es layers has been observed, the maximum single-hop ranges for F2 propagation being normally about 3,000 km, while, in addition, multi-hop echoes from ranges up to 12,000 km have been observed under favourable conditions. Information obtained with this equipment is found to be of considerable value in elucidating the various modes of propagation of radio waves over long distances.

The plan-position indicator in its present form is limited in angular resolution by the small aperture in wavelengths of the aerial but, nevertheless, provides a valuable method of observing the zones of the earth's surface to which radio waves are returned by the ionosphere. Such information could not be obtained by other means at present available. The area of the earth surveyed is normally 3,000 km in radius but can be very much greater under favourable conditions. A valuable addition to the p.p.i. is the speeded-up cinematograph facility, by means of which a vivid picture is obtained of the diurnal changes in the horizontal distribution of ionization in the ionospheric layers.

Acknowledgments

The authors wish to acknowledge the help and advice throughout the work of Mr. A. F. Wilkins, and the co-operation of Messrs. A. Wilson and E. Oakman in the mechanical design. They also wish to thank Mr. Caradoc Williams of the Royal Aircraft Establishment for making available the p.p.i. console.

The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of Radio Research of the Department of Scientific and Industrial Research.

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CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Constant-Frequency Oscillators

SIR,—I was interested to see Mr. A. S. Gladwin's article in *Wireless Engineer*, January 1956, in which the conditions are derived for the frequency of a regenerative oscillator to be independent of changes in the input and output resistances of the maintaining amplifier.

In this article, the harmonic frequencies of grid and anode currents have not been considered. I hope it will be of interest to compare the method developed by the author with the one which I described some time ago. This method takes into account the non-linearities of grid and anode circuits and, when the harmonic currents are not taken into consideration, the results are comparable with those obtained by Gladwin.

The general system reviewed is demonstrated in Fig. 1. The quadripole N is considered as being both linear and passive, hence its properties can be determined by means of equations

$$-V_g = Z_{gg} I_g + Z_{ag} I_a \quad .. \quad .. \quad .. \quad (1)$$
$$-V_a = Z_{ag} I_g + Z_{aa} I_a \quad .. \quad .. \quad .. \quad (2)$$

where all magnitudes V, I and $Z = R + jX$ are functions of the angular frequency ω.

The currents and voltages shown in Fig. 1 are, in these conditions, definite periodic functions of time. The electrical relation in the quadripole N will not vary if we replace the grid and anode circuits of the valve by generators of infinite internal resistance and identical current outputs as in the oscillator under consideration. The equivalent system so obtained is shown in Fig. 2. It is a linear system, to which the law of superposition can be applied.

Let us further assume the fact that the cathode current of the valve is a one-valued function of the equivalent potential on the grid surface.

$$i_c = i_a + i_g = f(v_e) = f\left(v_g + \frac{v_a}{\mu}\right)$$

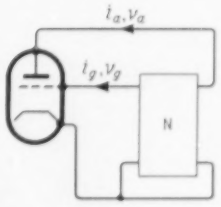

Fig. 1.

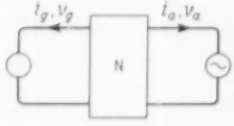

Fig. 2.

Having taken this into consideration and expressing voltages and currents in the form of harmonic series, we find—as we already know from Ref. 1.

$$\mathrm{Im}\left\{\sum_{k=1}^{\infty} k\left(V_{gk} + \frac{1}{\mu} V_{ak}\right) \cdot \left(I_{gk}^{*} + I_{ak}^{*}\right)\right\} = 0 \qquad (3)$$

where V_{gk} denotes the kth harmonic of grid voltage, etc.,

whereas I_{gk} and I_{ak} denote conjugate values up to $I_{g\infty}$ and $I_{a\infty}$.

Let us assume that I_{a1} is real and positive, and let us indicate $n_{ak} = I_{ak}/I_{a1}$, $n_{gn} = I_{gk}/I_{a1}$. We shall, by substituting (1) and (2) in formula (3), arrive at

$$\mathrm{Im}\left\{\sum_{k=1}^{\infty}\left(n_{ak}^{*} + n_{gk}^{n}\right)\left[\left(Z_{gg} + \frac{1}{\mu}Z_{ag}\right)_k n_{gk} + \left(Z_{ag} + \frac{1}{\mu}Z_{aa}\right)_k n_{ak}\right]\right\} = 0 \quad . . \quad . . \quad . . \quad (4)$$

where the index k next to the bracket indicates that the expression contained in the bracket should be considered as relevant to the kth harmonic. This formula permits us to expand the well-known Groszkowski formula; knowing the function $Z_{gg}(\omega)$, $Z_{ag}(\omega)$, $Z_{aa}(\omega)$ and the harmonic content of the grid and anode currents, it enables the frequency of oscillations ω to be determined.

In the case when the oscillator frequency depends chiefly upon the fundamental frequencies of the grid and anode currents, we may not take into account in formulae (4) the terms expressing the influence of higher harmonic frequencies. Providing that the value of n_{g1} is real, which is the case when the stability criterion (6) is fulfilled, and starting from Equ. (4), we arrive at

$$\left(X_{ag} + \frac{X_{aa}}{\mu}\right)_1 = -\left(X_{gg} + \frac{X_{ag}}{\mu}\right)_1 n_{g1} \quad . . \quad . . \quad (5)$$

It may be seen from this that, when the following conditions for the oscillator frequency are simultaneously fulfilled

$$X_{ag} = 0,\ X_{aa} = 0,\ X_{gg} = 0 \quad . . \quad . . \quad . . \quad (6)$$

the frequency is independent both of the amplifier's input resistance expressed by n_{g1} and of the valve μ. This criterion implies that the input, output and transfer impedances of the feedback network must be resistive at the oscillation frequency. This criterion has been related to the amplification factor μ of the valve instead of to its output resistance, because its value is less dependent on the external circuit and therefore it can be more easily determined.

In my paper[3] the reactances have been computed which, introduced into the grid and anode leads, fulfil the stability condition for the tuned-anode, tuned-grid, Colpitts, Clapp and Hartley oscillators. The fallacy of Llewellyn's formulae for a real circuit has been shown by an example in which the stabilizing reactance, calculated by means of these formulae, increases the dependence of frequency on the input resistance of the amplifier.

L. B. Lukaszewicz

Warsaw, Poland.

28th May 1956.

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3	+3
4	+5
5	+4
6	+4
7	+5
8	N.M.
9	N.M.
10	N.M.
11	N.M.
12	+4
13	+4
14	+4
15	+4
16	+5
17	+4
18	+5
19	+5
20	+5
21	+5
22	+5
23	+5
24	+5
25	+5
26	N.T.
27	N.T.
28	+5
29	+5
30	−3

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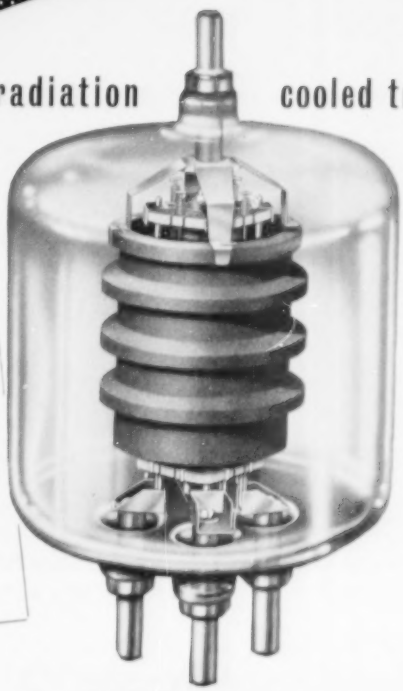

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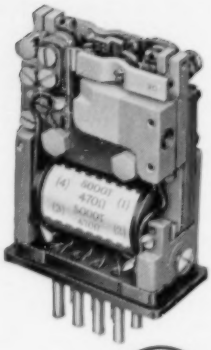

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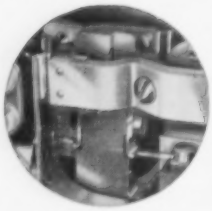

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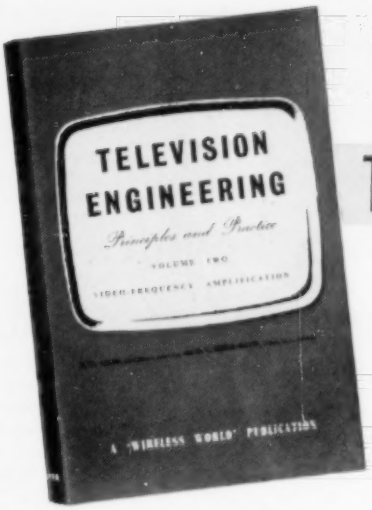

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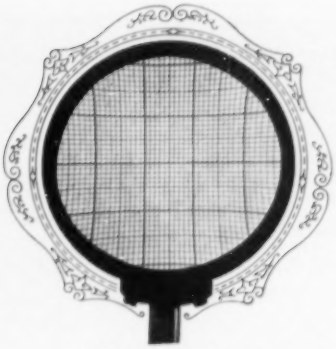

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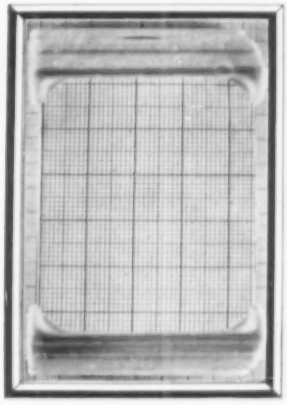

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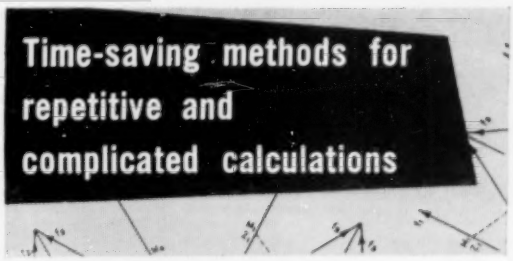

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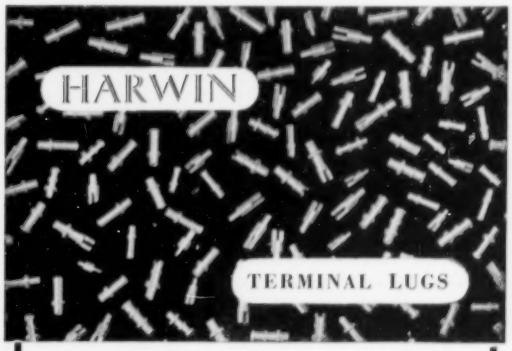

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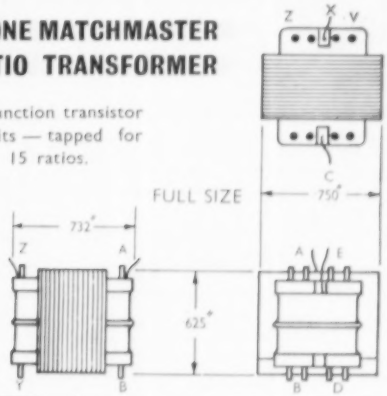

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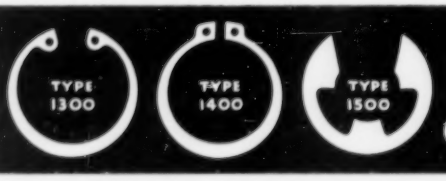

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